WIRELESS ENGINEER

Vol. XXIV. OCTOBER 1947 No. 289

EDITORIAL

Radiolympia

THE 15th National Radio Exhibition comes after a gap of eight years during which the energies of wireless engineers have been given chiefly to the development of apparatus for the purposes of war. It is natural, therefore, that the question of how wartime development has influenced the design of broadcast receivers should be prominent in one's mind, and that one should regard the exhibition as an opportunity for enquiry.

In the latter years of the war, when some of these developments had been partly revealed, it was fashionable in some circles to forecast enormous improvements in post-war receivers. Quite naturally this created the impression in the minds of the non-technical that these sets would be greatly superior in performance.

This view was never taken by those who understood the nature of wartime development, however, for although much of it has great peacetime application, comparatively little is of direct benefit to broadcasting as long as this is carried out on the present frequencies. The best-known developments of the war were in radar, pulse technique and centimetre-wave generators, radiators and feeders and are clearly unlikely to influence broadcasting much unless it moves to much higher frequencies than those now used.

The less publicized improvements of the war years, such as miniaturization and tropicalization, are of more benefit to the man-in-the-street. The former is most applicable to portable sets and hearing aids and the latter to equipment for overseas use. Neither is very spectacular, however. While a reduction of size and weight is obvious to the user, the benefits of tropicalization are apparent only in increased reliability in extremes of temperature and humidity.

At the time of writing only scanty details of the equipment shown are available, but they do indicate that these expectations are borne out. Most firms stress that their export models are designed for tropical use; there are quite a number of sets using some degree of miniaturization and all hearing aids are fully miniaturized. Television would seem to owe more to the war, for there are detailed improvements in circuits which can be traced back to the pulse technique of radar. In this radar is repaying the debt which it incurred to television in the early days, when much of its practice and many of the people concerned were drawn from television.

The "standard" broadcast receiver is still a four-valve three-band superheterodyne, however, with a triode-hexode frequency-changer, one i.f. stage, a duo-diode-triode providing detection, a.v.c. and a.f. amplifica-

tion, and a tetrode output valve. At first sight this would seem to owe little to the war years, but there are some indications of mechanical improvements. There are signs that the necessity for a robust, but light, construction to withstand the rigours of war has in some degree been carried over to peace.

A better judgment on these matters will be possible when the exhibits have been examined, and in the next issue of *Wireless Engineer* the apparatus at Radiolympia will be reviewed.

Ohm and his Law

FROM time to time articles are published in the technical press dealing with the question, What is Ohm's law? Such an article was published in the *Electrical Review* of 31st January and, as usual, led to some correspondence. In the same journal of 1st August a very interesting article by Mr. Thomas Carter dealt with the same topic. We were glad to see that he had been delving into the original treatises in order to find out exactly what Ohm did and what he said. It is an interesting pursuit, as we found in 1931, when preparing a paper on "The Nomenclature of the Fundamental Concepts of Electrical Engineering."* As Mr. Carter says, it is difficult to translate some of Ohm's terms because of the difference of atmosphere between then and now. He was writing before Faraday had discovered electromagnetic induction, when an electromotor was a device for imparting motion to electricity (that is, a primary battery) and when a "calorimotor" was merely a battery designed to give a large current and thus heat the wires. Even in 1862 the British Association committee refer to a "battery or rheomotor of unit electromotive force."

To Ohm's first paper on the subject in 1825 Poggendorff, the editor of the *Annalen der Physik*, added a footnote expressing the hope that Ohm would repeat his experiments with the thermo-electric circuit because the action would be steadier than with the hydro-electric circuit, in other words, that he should use thermo-junctions instead of batteries as the source of his current, a suggestion that sounds somewhat amusing at the present day, but which Ohm acted upon with great success. These examples show how careful one has to be in translating Ohm's descriptions of his experiments into modern phraseology. It must also be remembered that the idea of an electric current was relatively new, the electricity of the past having been almost entirely electrostatic and known to be a surface phenomenon. Ohm describes how he repeated an experiment made by Davy, who took a round wire and rolled it into a thin strip; when placed in a circuit the current was unchanged by substituting the strip for the round wire, although the surface was greatly increased, thus showing that the controlling factor was the cross-section and not the surface, and that the electric current flowed through the mass of the conductor as heat flows from a hot to a cold part of the material.

* *J. Instn. elect. Engrs.*, Vol. 70, No. 420, Dec. 1931. p. 54.

Georg Simon Ohm

Before discussing his experiments, however, let us look at the man himself. Georg Simon Ohm was born in Erlangen in Bavaria on 16th March, 1789, and not in 1781, as stated by Mr. Carter, nor in 1787, as stated in several encyclopædias, including the "Britannica." He died in Munich on 7th July, 1854. His father was a locksmith but took a keen interest in philosophy and mathematics, and helped and encouraged his sons in their mathematical studies. On leaving school he went to Erlangen University, but after three terms went to Switzerland as a teacher in a private school, where he remained for three and a half years, then going to Neuenburg in Prussia as a mathematics teacher. He continued his studies during this period, and in 1811, after six months in Erlangen, he was awarded the Ph.D. degree. He was recognised by the University as a coach in mathematics, but again after three terms he left the University and was appointed a teacher at a school in Bamberg, where he remained until 1817, when he went to Cologne as a senior teacher of mathematics and physics at the gymnasium or high school, formerly the Jesuit College.

It was here that he began his experimental researches and published his first papers on

the subject. There were no facilities for his experiments in the School and they were carried out in his lodgings in his spare time, of which he had so little that he applied for leave of absence for the school year 1826–1827 in order to concentrate on his research. This was granted at half salary, but at the end of the period he was so dissatisfied with the conditions under which he would have to work if he returned, that he resigned and went to Berlin, where he gave lectures on mathematics at the Military School. It was a financial sacrifice, but it gave him the leisure necessary to continue his experimental researches. He had a brother Martin Ohm, three years younger than himself, who became Professor of Mathematics at Berlin University and published many papers. He also lectured at the Military School and as he had been in Berlin since 1821, one can understand why Georg Ohm was attracted to Berlin.

After six years in Berlin, Ohm went to Nuremberg in 1833 as professor of physics in the Polytechnic of which from 1839 to 1849 he was the Director. Here again he found that his duties allowed too little time for research and in 1842 he asked for and obtained an assistant to help in the administration. What was almost certainly the greatest event in his life occurred in 1841 when the Royal Society awarded him the Copley medal in recognition of his outstanding achievements ; this was followed in 1842 by his appointment as a foreign member of the Society. These distinctions must have been very gratifying to one who undoubtedly felt that his scientific work was not fully appreciated by German scientists and who had reached the age of 52 without achieving his ambition of being appointed to a University Chair. This he achieved in 1849 when he was appointed Professor of Physics in the University of Munich ; he also acted as consultant to the Bavarian telegraph authorities and as Conservator of the Bavarian State Mathematical and Physical Collection, but his tenure of these offices was short for he died in 1854.

Even if his work did not receive the immediate general recognition that he expected, there were a number of German physicists who saw the importance of it ; for example, Fechner said in 1929 that he must give Ohm the credit for having with the few letter symbols of a simple formula laid the foundations of a new epoch in the teaching of galvanic electricity.

As an indication of the conditions under which his experimental work was carried out he mentions in a communication to Schweigger's Journal in 1827 that he had discovered that some discrepancies in a previous communication were due to temperature changes in the attic-kitchen in which the experiments had to be carried out. This was in Berlin just after he had left Cologne.

He had only been dead seven years when at the British Association meeting in Manchester in 1861, Charles Bright and Latimer Clerk proposed that the unit of electromotive force should be called the ohmad—presumably to rhyme with farad—but in the report of the committee in 1864 the name " ohmad " is given to the B.A. unit of resistance.

It is not generally known in this country, or even in Germany, that the Ohm was a unit of liquid measure long before it had any electrical significance. It was used not only in Germany but in Scandinavia, Holland and Switzerland, mainly as a wine measure ; it differed slightly in different countries, but was about 40 gallons.

First Statement of the Law

Turning now to Ohm's publications, the first, except for a purely geometrical paper, appeared in the *Annalen* in 1825 ; it was entitled " Preliminary Intimation of the Law according to which Metals Conduct Contact Electricity." He explained that it was very incomplete but that he had little leisure for prosecuting the experiments. As a matter of fact, his law was a logarithmic one and quite wrong, mainly because of the fluctuation of current due to polarization of the battery ; hence Poggendorff's suggestion that he should use thermo-junctions instead of batteries.

A much more important paper followed in 1826; it was entitled " Determination of the Law according to which Metals Conduct Contact-electricity, together with an Outline of a Theory of the Voltaic Apparatus and the Schweigger Multiplier." He describes fully the thermo-junctions of bismuth and copper, one in boiling water and the other in ice, with which the current was generated, and the current measuring apparatus consisting of a magnetic needle suspended over a horizontal conductor by a thin strip of gold from a graduated head. He states that the current is given by the formula $a/(b+x)$ where a is proportional to the e.m.f. (erregen-

den Kraft), b is the internal resistance and x the resistance (Widerstandslänge) of the external conductor. He uses the term resistance-length because, having no unit of resistance he had to express it as the equivalent length of a certain wire.

In view of what has been said about Ohm's reluctance to use the term "resistance," it is interesting to note that in this very early paper, the word "Widerstand" occurs once, and "Widerstandslänge" six times on a single page; in some later papers he preferred "equivalent length." He confirmed the law that cylindrical conductors of the same kind but of different diameters have the same conductance if their lengths are proportional to their cross-sections. He says that Barlow and Becquerel had found the same result. Later in the same year (1826) he published a paper entitled "An Experimental Theory of the Electroscopic Phenomena Produced by Galvanic Forces." The electroscope was the instrument used to measure potential, and electroscopic phenomena were voltage phenomena. This is a very important paper. He gives two formulae; viz. (in modern symbols),

$$\text{(a) } I = \sigma a \frac{V}{l}, \text{ (b) } V_x - V_o = \pm \frac{x}{l} V$$

in which σ is the conductivity, a the cross-section, l the length, V the potential difference (Spannung) between the ends, and V_x the potential at distance x along the conductor from the point at which it is V_o. He states then two laws that follow from formula (a);

I(a). The current is unchanged if, with the same V and σ, the length is made proportional to the cross-section.

I(b). The current is unchanged if, with the same V and a, the length is made proportional to the conductivity.

II. If σ and a are unchanged the current is proportional to V/l.

He also says that (a) may be written $I = V/l$ if by l one understands the equivalent and not the actual length.

He again disclaims any originality for law I(a); he says that it was discovered by Davy and confirmed later by Barlow, Becquerel and himself. The remainder of the paper is devoted to a study of the voltage drop around a circuit and the potential of any part of it.

It will thus be seen that Ohm's law, as generally understood, was published with experimental verification in 1826 that is, the year before the publication in Berlin of his paper entitled "Die Galvanische Kette, Mathematisch Bearbeitet" (The galvanic circuit mathematically treated). This was a much larger publication and is the only one of his many papers that the "Encyclopaedia Britannica" regards as of the first order. It deals with the potential gradient and its relation to the other properties of the circuit, with conductors in series and in parallel and many other aspects of the galvanic circuit. What we call potential difference Ohm calls Elektrische Spannung or Differenz der Körper. Although it is more mathematical and discusses the problems at greater length, it adds nothing of fundamental importance to the statements of the preceding paper regarding Ohm's law.

In a paper communicated to *Schweigger's Journal* in 1829 Ohm says that the most general result of the experiments on the nature of the conduction of electricity is that the current in a galvanic circuit is (a) directly proportional to the sum of all the e.m.fs. in the circuit, and (b) inversely proportional to the sum of all the resistances. In this paper he discusses the relative merits of the terms resistance and equivalent length and expresses his preference for the latter; he says that the word resistance gives a very imperfect picture of what is happening. Although during the following two or three years he contributed several lengthy papers on electrical matters, they do not contain anything that could affect one's views on Ohm's law. Most of his subsequent publications dealt with acoustics and the polarization of light, his last paper (1853) of nearly 200 pages being devoted to the latter subject and optical interference phenomena.

What is Ohm's Law?

We must now return to our opening question; What is Ohm's law? Is one justified in saying that certain conducting materials do not obey Ohm's law and can be referred to as non-ohmic materials? The first law that Ohm states definitely as a law and prints in italics, while disclaiming originality for it, is that cylindrical conductors of the same kind and different diameters have the same conductance if their lengths are proportional to their cross-sections. This would not apply to a non-ohmic material such as silicon carbide since on replacing one wire by another of smaller section the

current density would be different and therefore also its resistivity, and the length would have to be modified accordingly.

Instead of non-ohmic some people prefer non-linear, referring, of course, to its characteristic and not to the resistor itself, since a non-linear resistor may be a straight rod and a linear one a helically wound wire.

When formulating his law Ohm undoubtedly regarded the conductivity of the material as a constant and, apart from any secondary effect due to change of temperature, independent of the current flowing through it. When Ohm wrote the formula $I = \sigma a V/l$, he certainly expected I to be proportional to V in any given conductor; he knew of no such substances as silicon carbide. Of course, whatever tricks the material may play, one can always find a value of σ to satisfy the equation $\sigma = lI/aV$, but is this a law or mere lawlessness? Is it not inherent in Ohm's law that, apart from any secondary effects, the current is proportional to the voltage; that is, that σ is constant and not a variable like the permeability of iron? If so, then one is perfectly justified in referring to a material in which σ varies with the current as not conforming to Ohm's law. It seems very probable that materials of the silicon carbide type will come into more general use and it is necessary to have some simple designation so as to distinguish between them and ordinary resistances. Non-ohmic seems quite harmless, but as an alternative we may go back to the eighteen-sixties and say non-ohmad-ic, which would probably be shortened to nomadic,—a very descriptive designation for a resistance in which the value of the conductivity wanders about with every change in the current.

G. W. O. H.

OSCILLOSCOPE TIME-BASE CIRCUIT*

by H. den Hartog and F. A. Muller

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SUMMARY.—This paper is an analysis of the mechanism of synchronization of conventional time-base generators, and a description of an improved circuit having an extended range of synchronization and automatic indication of the time-sweep frequency.

TIME-BASE circuits of the Puckle type have excellent linearity, high stability of operation, good flyback ratio, and a high upper frequency limit, but have certain drawbacks which detract from their utility when synchronization is important.

The analysis of saw-tooth generators, from the point of view of synchronization is best considered with the aid of the circuit diagram in Fig. 1. The capacitor C is charged by the constant current I_0 until the cathode of valve B reaches the threshold voltage at which it begins to conduct. Then, by trigger action, C is rapidly discharged through valve B; this completes the cycle and restores the original situation.

It is shown in Fig. 2 how the voltage V_x between the cathode of valve B and earth drops in a uniform fashion until the threshold voltage is reached. (For simplicity the flyback time has been neglected. The natural frequency $f_0 = 1/T_0$ of the saw-tooth oscillation is determined by the charging current I_0, the capacitance C, and the distance V_0 between the limiting voltages of the saw-tooth, by the equation

$$f_0 = \frac{I_0}{C V_0} \quad .. \quad .. \quad .. \quad .. \quad (1)$$

Thus the natural frequency f_0 of the saw-tooth oscillation can be adjusted to any desired value by suitably adjusting I_0, to which it is directly proportional.

Now the threshold voltage is determined by the grid potential of valve B, and the usual method of synchronizing the time sweep with an external signal is based on superimposing the signal on this grid potential.

Again for simplicity we shall suppose that the synchronizing voltage is impressed between the anode of valve A and the grid of valve B, as shown in Fig. 1, and for the same reason we shall suppose this voltage to be sinusoidal and of peak value V_s.

The line representing the threshold voltage

* MS. accepted by the Editor, October 1946.

in Fig. 2 is no longer straight but undulating, as indicated in Fig. 3. The phenomenon of synchronization is explained by the fact that the distance between the voltages limiting the saw-tooth wave is no longer constant, but varies between $V_o + V_s$ and $V_o - V_s$, so that, depending on the frequency of V_s and the resulting phase relationship between it and the saw-tooth wave, the saw-tooth frequency may vary between $I_o/C\ (V_o + V_s)$ and $I_o/C\ (V_o - V_s)$.

By this action the frequency of the time

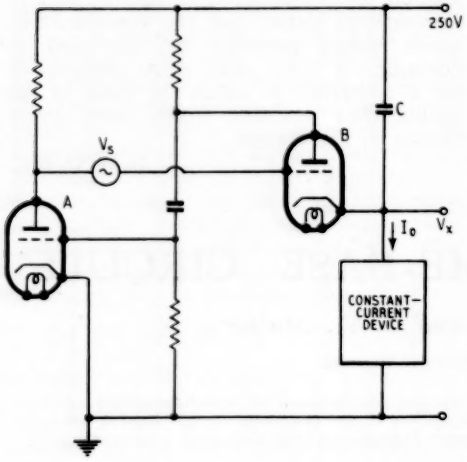


Fig. 1. *Fundamental Puckle time-base circuit.*

sweep may be lowered or raised (with I_o kept at a constant value), but at the same time the amplitude varies inversely as the frequency, the maximum relative variation being V_s/V_o in both amplitude and time, and approximately also in frequency:

$$\left(\frac{\Delta f}{f_o}\right)_{max} \approx \frac{V_s}{V_o} \qquad .. \quad .. \quad .. \quad (2)$$

Now the frequency of the saw-tooth oscillation cannot be lowered indefinitely by increasing V_s, as might be inferred from Equ. (2), since the additional condition has to be fulfilled that the sloping part must have only one point of intersection with the sinusoidal threshold voltage.

In order to make this point clear, let us assume that starting from Fig. 3(b) an attempt is made to lower the synchronized saw-tooth frequency still further. It is clear that this cannot possibly be done without increasing V_s, as the largest amplitude attainable is $V_o + V_s$, and in consequence the lowest frequency will be $I_o/C\ (V_o + V_s)$, which is the limiting case illustrated by Fig. 3(b).

The lower limit of the range of synchroniza-

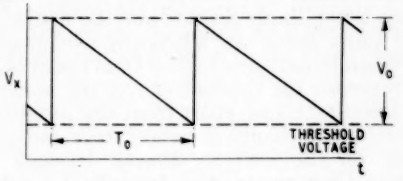


Fig. 2. *Output voltage of the circuit of Fig. 1.*

tion cannot be further reduced by increasing V_s, as might at first be expected, because the minima of the sinusoidal threshold can no longer be reached by the sloping part of the saw-tooth without a previous intersection. What actually happens on increasing V_s and lowering the frequency of the synchronizing signal is shown in Fig. 4. The result is distortion of the time sweep and a confused pattern on the oscilloscope screen as shown in Fig. 5.

For this reason the limit of smooth control of the saw-tooth frequency by synchronization is reached when, at the danger point indicated in Fig. 3(b) by an arrow, the sloping part of the saw-tooth is a tangent to the sine wave.

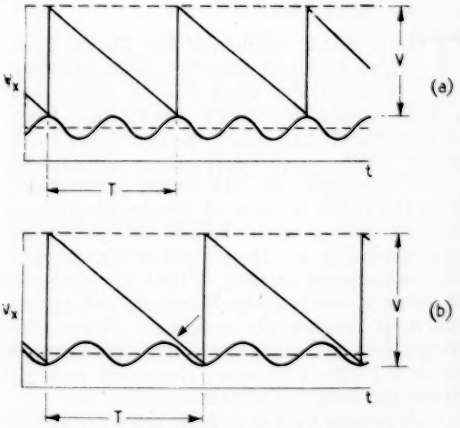


Fig. 3. *Output voltages of the circuit of Fig. 1, with a sine-wave synchronizing voltage applied.*

It may be shown from geometrical considerations that V_s must be kept below the limit given by

$$V_s < \frac{V_o}{4{\cdot}55n - 1} \quad .. \quad .. \quad .. \quad (3)$$

for synchronization with the frequency of the nth harmonic. From this result it follows that the maximum relative frequency decrease through synchronization with n sine-waves per saw-tooth cycle is

$$\left(\frac{\Delta f}{f_o}\right)_{\text{max sync}} = \frac{1}{4{\cdot}55n} \quad .. \quad .. \quad (4)$$

or 22, 11, and 7 per cent for the 1st, 2nd, and 3rd harmonics respectively.

The higher-frequency limit of the range of synchronization, can in fact be raised by increasing V_s, but as soon as Equ. (3) is violated it will be effected only at the expense of the lower-frequency limit. The entire

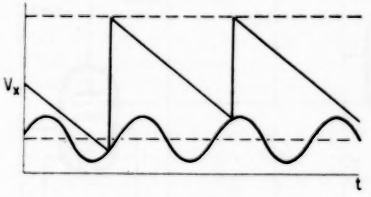


Fig. 4. *Distorted saw-tooth oscillation resulting from too large a frequency shift in the synchronizing signal.*

range is thus moved up the frequency scale even to the extent of excluding the natural frequency f_o of the time sweep. Besides the possibility of patterns like the one shown in Fig. 5, the general result is a marked reduction of the sweep amplitude.

As it is wise to keep well below the danger limit, it follows that only frequency deviations of the order of 10 per cent can be comfortably used.

In case of large frequency shifts in the signal under observation it becomes necessary to readjust the time-base oscillator by hand by increasing or decreasing I_o according to whether the shift is towards higher or lower frequencies.

To sum up; the inherent limitation of conventional synchronization systems lies in the fact that it is effected, with constant charging current, through varying the amplitude of the saw-tooth wave by plus or minus V_s, and that this variation must be kept comparatively small to prevent erratic behaviour.

Extension of Synchronization Range

It would obviously be much better if the charging current I and not the amplitude V of the saw-tooth wave could be changed, for there are no such narrow limits for the admissible variations of I as for V. This leads one to investigate the following line of attack :—

A synchronizing voltage V_s with a frequency lower than the natural frequency of the time base will increase the sweep amplitude V, as was shown in the preceding section and more especially in Fig. 3(b).

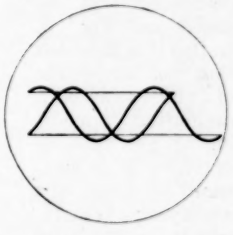


Fig. 5. *Pattern appearing on the oscilloscope screen when the oscillation of Fig. 4 is used as a time base.*

Now if this increase is itself made to cut down I, it will automatically reduce the natural frequency of the time base, much as one should do by hand when warned by the increase of the sweep amplitude that the point where the synchronizing signal will lose control is being approached. In this

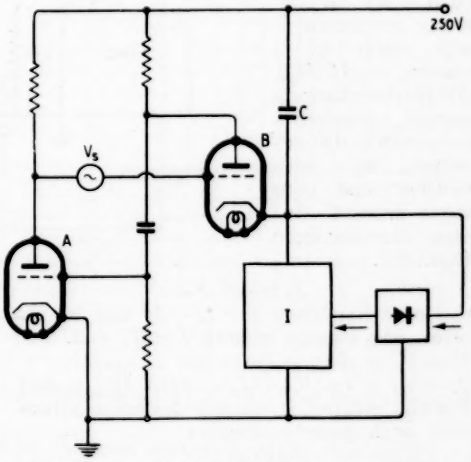


Fig. 6. *Fundamental circuit for automatic charging-current control.*

way, the natural frequency will come nearly to coincidence with the synchronizing frequency, and because of this reduced frequency shift the resulting change in amplitude will be quite small.

It will be noted that this mechanism may

be described as an inverse feedback applied to the amplitude, but the remarkable feature of this feedback is that it is only operative when a synchronizing voltage is present. Without a synchronizing voltage the charging current I has no influence on the amplitude V.

This system of extended synchronization will be indicated by the term *panchronization*, and the discussion will be based on Fig. 6. In this it is assumed that the amplitude of the saw-tooth oscillation appearing at the cathode of valve B is measured by means of a rectifying circuit, which in its turn feeds a control voltage to the constant-current device (indicated by I) which is thereby controlled in such a way that a change ΔV in the sweep amplitude V will result in a change $-\Delta I = g.\Delta V$ in the charging current, g being a constant determined by both rectifier and constant-current device. Starting from Equ. (1):

$$f_o = I/C\,V_o$$

the new frequency $f = f_o - \Delta f$ will be the result of a change in both $I = I_o + \Delta I$ and $V = V_o + \Delta V$, so that

$$f_o - \Delta f = I/C\,V = (I_o + \Delta I)/C\,(V_o + \Delta V)$$

For the greatest frequency deviation attainable with panchronization

$$\Delta V = V_s,$$

(as is the case with synchronization)

and accordingly

$$\Delta I = -gV_s$$

Eliminating from these four equations I_0, I, and V, we obtain

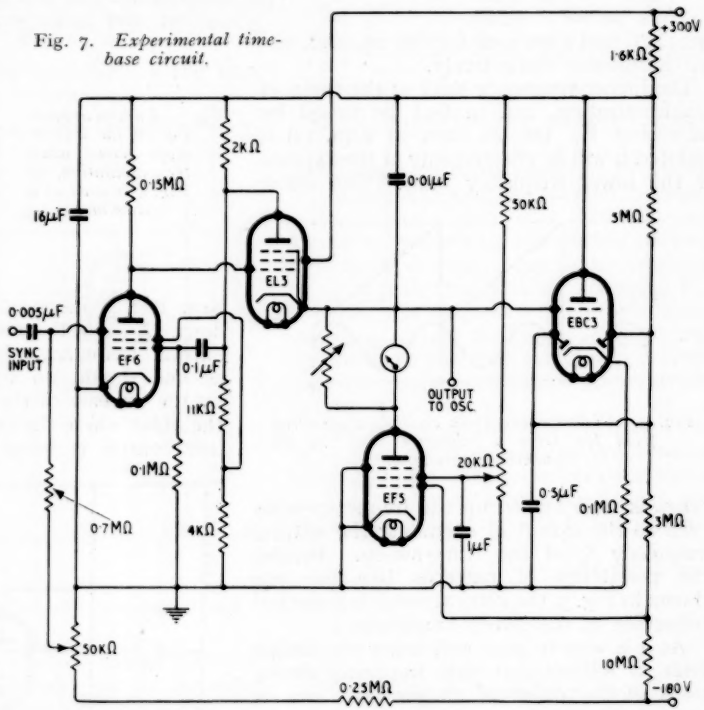
$$\left(\frac{\Delta f}{f_s}\right)_{max} = 1 - \frac{V_0}{V_0 + V_s}\left(1 - \frac{g}{I_0}V_s\right),$$

and when we consider that V_s is quite small, we obtain the approximate expression

$$\left(\frac{\Delta f}{f_0}\right)_{max} \approx \frac{g}{I_0}V_s \quad .. \quad .. \quad .. \quad (5)$$

as might be expected, gV_s/I_0 being the relative change in charging current, and the direct effect of the amplitude variation on the range of panchronization being negligible.

Fig. 7. *Experimental time-base circuit.*



Comparing Equs. (5) and (2) it is found that with the same very small signal V_s, panchronization is gV_0/I_0 times more effective than synchronization.

For a pentode valve as the constant-current device, g/I_0 will be about 0.5 per volt (assuming some loss in the connecting network), so that with $V = 200$ volts the superiority of panchronization is represented by a factor of 100.

The Practical Circuit

In order to try out this system, the experimental circuit shown in Fig. 7 was adopted. In addition to the panchronizing circuit, some deviations from usual practice were included

that are also useful with the ordinary synchronizing circuit. Among these are :—

(1) a lowering of the anode-voltage supply of valve EL3 with respect to the screen voltage, resulting in anode-current cut-off before the cathode current drops to zero, and an accordingly improved flyback ratio ;

(2) a negative bias on the suppressor grid of valve EF6. (This bias, besides providing an amplitude control, is useful in preventing grid rectification of the synchronizing signal from causing a steady potential to be superimposed on the threshold voltage which would make synchronization possible only with frequencies above the natural frequency.)

The operation of the panchronizing circuit is as follows. In order to avoid distortion, the saw-tooth oscillation appearing at the cathode of valve EL3 is applied to the control grid of the EBC3 cathode-follower valve, so that it again appears at the cathode of this valve but at a much lower impedance level.

Now by means of the diodes in the EBC3 valve a rectified voltage is made to appear across the 0.5-μF capacitor in such a way that an increase in the sweep amplitude results in a decrease of the steady voltage across it. By means of a potential-divider connection this voltage is fed to the EF5 constant-current pentode. In this fashion the anode current of this pentode is decreased by an increasing sweep amplitude, thus lowering the frequency of the time base in accordance with the lower frequency of the signal that is the primary cause of the rise in sweep amplitude.[1]

With a signal of 1.5 volts r.m.s., the range of panchronization of this circuit is from 150 to 1,700 c/s, the amplitude, as measured on the oscilloscope screen, varying from 67 to 58 mm and the charging current varying from 0.36 to 3.5 mA. With only 0.45 volts r.m.s., the range is still remarkable : 250 to 650 c/s, with the amplitude varying from 67 to 65 mm and the charging current varying from 0.6 to 1.5 mA.

These figures refer to panchronization with the fundamental frequency ; i.e., with one sine-wave per saw-tooth cycle.

When the frequency of the signal V_s is increased above the limit of panchronization in the fundamental frequency, the circuit jumps over to panchronization in a higher harmonic without losing control, and the time-base frequency continues locked to that of V_s.

In a certain case, one sine-wave was obtained on the oscillogram from 120 to 1,900 c/s, two waves from 210 to 3,740 c/s, and three waves from 300 to 5,620 c/s, the r.m.s. panchronizing voltage being 1.5 volts. Even with as little as 0.5 volt there was ample overlapping of these ranges : one wave from 300 to 1,000 c/s, two waves from 640 to 1,900 c/s, and three waves from 970 to 2,700 c/s. (The number of waves for a certain signal frequency could be influenced by manipulating the suppressor-grid bias control if desired. Because of the amplitude-frequency feedback the main influence of this bias is on frequency rather than on amplitude as is the case with synchronization.)

Frequency Metering

As the amplitude of the time base is kept substantially constant, the charging current is almost proportional to the frequency. This is borne out by the graph of Fig. 8.

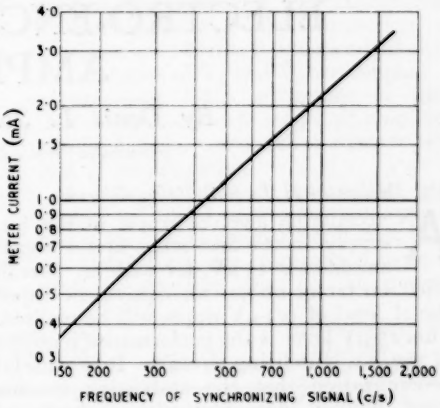


Fig. 8. *Dependence of meter current on frequency.*

This feature makes possible the incorporation of a frequency meter of fair accuracy in the time-base circuit. It is best to use a logarithmic milliammeter, calibrated in such a way that one has to multiply its reading by the frequency indicated on the time-base selector switch (selecting capacitor C) in order to find the frequency of the signal under observation.

The adjustment of the circuit is as follows:

[1] As the saw-tooth oscillation shows self-rectification (as is evident from Fig. 3) a separate rectifier is not an absolute necessity. The EBC3 cathode-follower may also be used as a phase-inverter to obtain push-pull output for oscilloscopes requiring symmetrical deflection.

The time-base selector switch is turned to the 50-c/s (power-line frequency) position, and a 50-c/s signal is impressed. Then the meter shunt resistance is adjusted until the meter points to the scale division marked 1. The signal is then removed and the screen-grid voltage of the EF5 charging pentode is adjusted until the meter points to 1 again. (To check the accuracy of adjustment, this procedure may be repeated.) In this way the natural frequency of the circuit is made to coincide with the frequency-meter mid-point, thus providing an easy means of reproducing one definite setting of the circuit at which the calibration has been carried out.[2] These adjustments need only be made occasionally, the two potential dividers being installed behind the panel and only accessible with a screwdriver.

The operating frequency is controlled by the EF6 suppressor-grid bias control, this and the panchronization attenuator (not shown in Fig. 5) being the only knobs that may (but hardly need) be manipulated.

The addition of panchronization and automatic-frequency indication makes the oscilloscope suitable for many new and unusual applications. However, it takes some time to get used to it, as it is at first strange to discern on the screen only changes in the shape of observed signals, and to have to depend on the frequency meter for changes in frequency.

[2] In doing this the natural amplitude should also be adjusted to a definite value by means of the EF6 suppressor-grid bias control, but in practice this has little effect on the calibration as the trans-conductance of the EF5 valve is nearly proportional to anode current within wide limits, and besides this only affects the small departure from proportionality of meter current to frequency shown in Fig. 8.

Acknowledgments

The authors wish to accord their thanks to Prof. J. Clay for the interest taken in their experiments on c.r. oscilloscope construction, and to Sgt. Maurice Bierman of the Royal Dutch Meteorological Institute, De Bilt, for developmental work done on this time-base circuit.

ELECTRO-ENCEPHALOGRAPH AMPLIFIER

By Denis L. Johnston, B.Sc.(Eng.)

(*Concluded from page 277, September issue*)

3.3. *Design of H.T. Stabilizing Circuit*

An acceptable level of ripple in the h.t. supply to the first stage of the signal amplifier is 15 μV, for with an overall rejection factor of 50, an equivalent input signal level of 0.3-μV ripple will be realised. This 15-μV level is the performance required of the h.t. stabilizing circuits. It is a useful generalization that the stabilizing circuits must employ degeneratively a degree of amplification of about one order less than the degree of amplification of the signal-amplifier itself.

Consequently we can anticipate that the stabilizing circuits will contain roughly as many stages as the amplifier circuit. In a multi-channel amplifier it is possible to economize by using one power supply for several channels, and in a four-channel equipment it is convenient to use two two-channel amplifier units with two associated power units. The complete circuit, Fig. 31, shows only one of two identical amplifier channels.

An electronically-stabilized supply used "floating" to earth must have no effective leakage or there will be trouble with ripple voltages in the high-impedance grid-input circuits of the signal amplifier. A high degree of insulation can be provided in mains-driven rectifier-stabilizer circuits, but it is an uneconomic way of attaining the required low level of leakage. It is easier to design with normal insulation and screening and then control the effect of leakage current by degenerative stabilization with respect to earth.

It can be shown that in a differential amplifier the overall stabilization of h.t. should be of a higher order than the stabilization of the h.t. to earth. Accordingly, it is better to stabilize the h.t. overall as in Fig. 29(*a*), and then control the potential to earth with an auxiliary stabilizer, than to use

one stabilizer between earth and positive and another between earth and negative as in Fig. 29(*b*).

The output impedances obtained in the present arrangement are tabulated in Fig. 28.

If all the amplification applied degeneratively over the two stages of the stabilizer is used in one stage the same degree of stabilization is obtained at an output impedance of less than one ohm. But there is a tendency towards instability at a fairly high audio frequency, so it is not used in this case. The output impedances tabulated in Fig. 28 are adequate because the load current is constant and the amplifiers push-pull. The overall arrangement of the h.t. stabilizer and the control circuits is set out in the schematic circuit Fig. 30.

The detailed circuit of the h.t. stabilizer appears at the left hand side of Fig. 31. The division between the power unit and the amplifier chassis is made at the point in the circuit indicated by the dotted line.

A reference voltage of 30 V is derived from a potential divider connected across the voltage stabilizer valve V_3, and is applied through an RC filter of about 10-seconds time constant to one grid of the amplifiers $V_{1.2}$ and $V_{6.7}$. The other grid of $V_{1.2}$ is connected to a potential divider across the 220-volt output of the stabilizer circuit. The difference voltage is amplified and applied to V_4 which controls the current in the 5-kΩ series load.

The second grid of amplifier $V_{6.7}$ is connected to a potential divider across the "−70 to E" output point. The difference between this potential and the reference voltage is amplified and applied to V_5 which controls the current in the cathode-coupling load of V_4 and V_5. The two sections of the stabilizer are effectively independent although the cathodes of V_4 and V_5 are coupled. This arrangement takes advantage of the fact that coupled-cathode stages of amplification can be directly connected in cascade without loss of amplification.

A potential divider across the 320-volt supply is tapped at a point proportional to 220 volts. This potential and that from the 220-volt supply are transmitted back to the grids of amplifier $V_{8.9}$ on the power unit

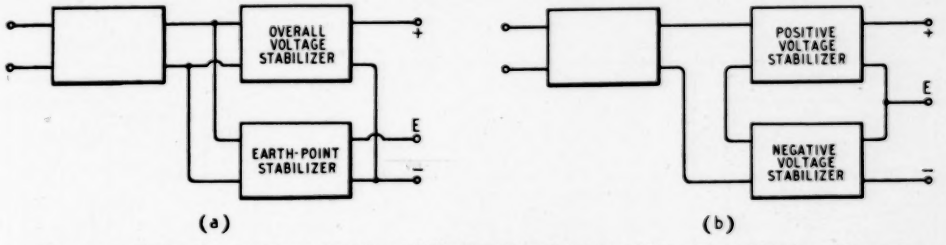


Fig. 29. *Schematic arrangement of the stabilized earth point system (a) and combined positive and negative voltage-stabilizing circuits (b).*

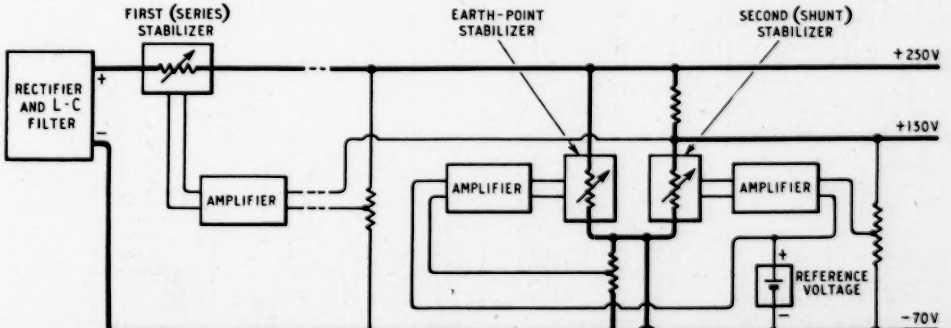
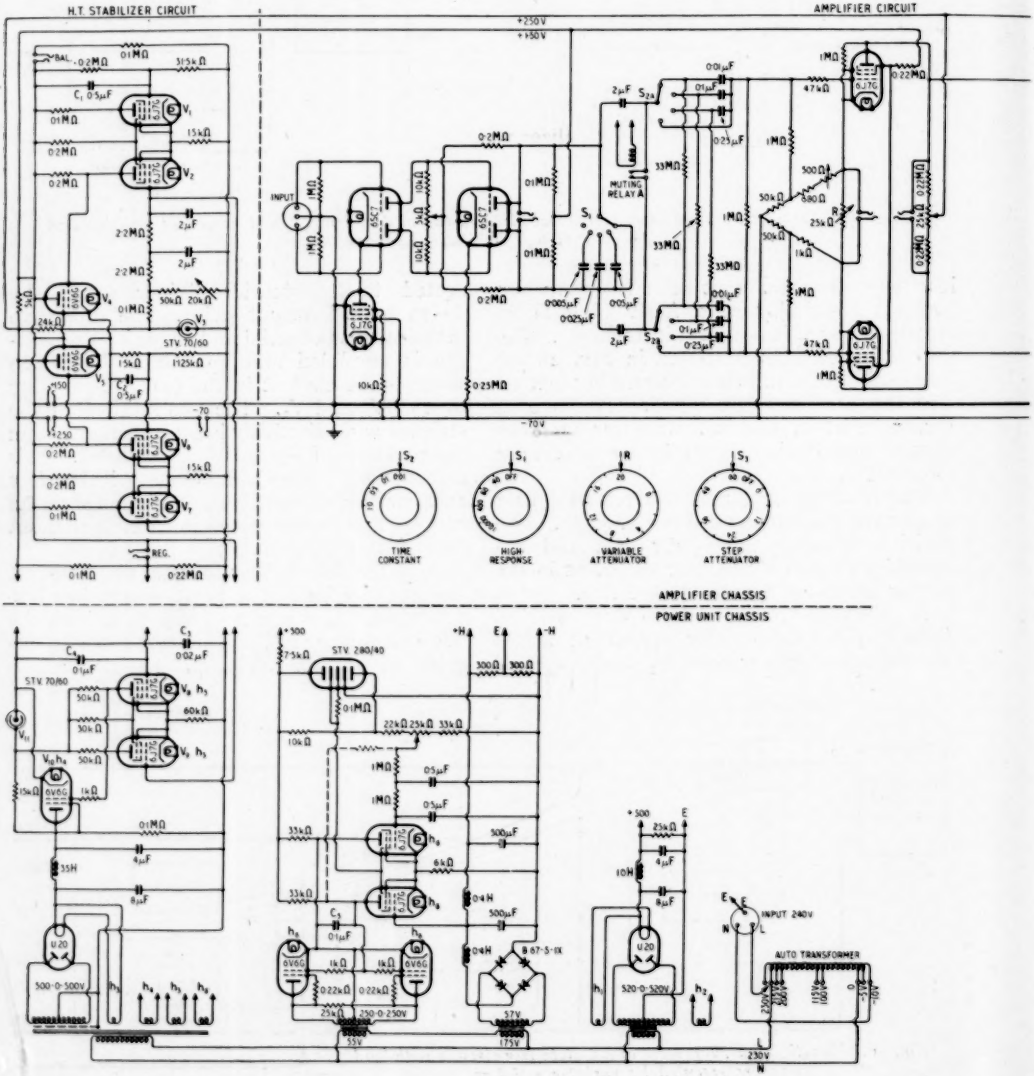


Fig. 30. *Functional arrangement of h.t. stabilizer circuits with load valves represented by variable resistors.*

chassis. The amplified difference voltage is applied to the grid of the series load valve V_{10}. The cold-cathode valve V_{11} relieves part of the load on V_{10} as the limit of anode dissipation is approached, thus extending the range of working voltage, which for this first stage of stabilization is represented by Fig. 32. The figures of overall performance have already been given.

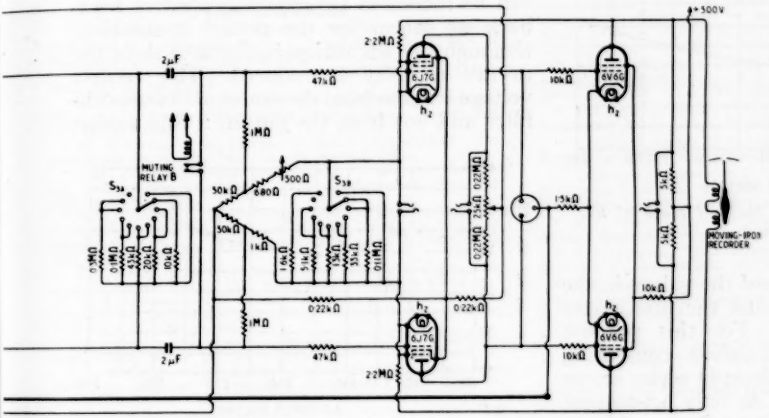
The potential dividers are adjusted to exact values of resistance. Samples of the valves used as V_3 vary widely from the nominal value of 70 volts, so the 20-kΩ variable resistor is provided for setting up the correct voltage. All resistors in the potential-divider circuits are wire wound for the sake of stability.

The purpose of the capacitors C_2, C_3, and C_5, is to suppress spurious oscillation at supersonic frequencies. The values indicated are about ten times the minimum values necessary to achieve stability in order to



confer a margin of safety. If the value is excessive these capacitors depreciate the stabilization to sudden changes of mains voltage. On the other hand, C_1 and C_4 improve such stabilization and reduce mains-frequency ripple.

There follows a summary of the arrangements that have been discussed and are employed to effect stabilization of high-tension supplies down to unusually low levels:



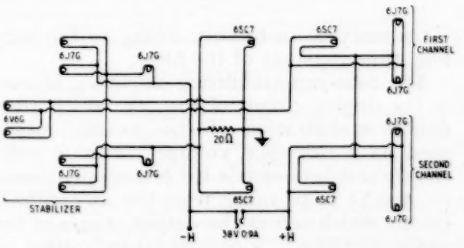


Fig. 31. *Circuit of amplifier and stabilized power supplies with the series-parallel connection of valve-heaters to d.c. stabilized supply shown separately. Only one amplifier channel is included—the other being identical.*

(*a*) Reference voltage derived from cold-cathode valve, with smoothing filter to eliminate fluctuations.

(*b*) Degenerative stabilizing circuits used throughout to avoid the inconvenience of balancing adjustments.

(*c*) Coupled-cathode differential amplifier circuits used to facilitate direct interstage coupling and reduce effect of fluctuations in heater supply.

(*d*) Provision of stabilized direct-current heater supplies where necessary.

(*e*) Easing of insulation and screening requirements by use of "floating" h.t. supply regulated with respect to earth potential.

(*f*) Differential cancellation of induction and voltage drop in connections between power-unit chassis and amplifier chassis.

(*g*) Division of stabilizing circuits between power-unit and amplifier chassis.

3.4. *Power-unit Chassis*

The transformer and chokes of the power unit cannot be far removed from the amplifier chassis in a self-contained equipment. The actual magnetic field experienced at the top of the amplifier chassis at the input stage is 2 gauss, and on the underside 0.3 gauss. The ripple picked up in the amplifier can only just be detected when a recorder is used at maximum sensitivity.

As a precaution the smoothing chokes are enclosed in boxes of 18 s.w.g. mild steel. The transformers are designed for only 75 per cent of the normal flux density of 12,000 gauss. Construction of the power-unit chassis is such that it can be cased-in completely with mumetal, but this is not necessary. All transformers are wound for 230-volt supplies and an auto-transformer is fitted in the base of the cabinet to provide for the range 100-255 volts. There is an appreciable electric field from the anodes of the rectifier valves, and consequently the stabilizer and early amplifier 6J7G valves are fitted in screening cans, and the grid leads screened.

Smoothing of the rectified h.t. current is accomplished by the stabilizer circuit more readily than with RC filters. The filtering

shown in Fig. 31 is more than adequate, only a single choke being used. In a practical case where two chokes were used in a smoothing filter the ripple was five times the calculated value due to induction between the chokes although they were fairly well spaced.

3.5. *Stabilized Heater Supply*

Adequately regulated direct current is

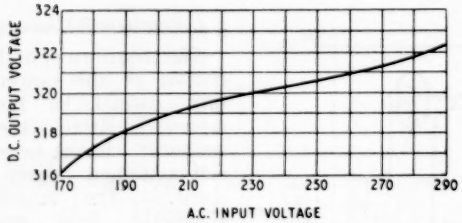


Fig. 32. *Performance of the first stage of h.t. stabilization.*

necessary for the heaters of the valves in the final stabilizing circuits and the first stages of signal amplification. For this purpose Miller[39] used 12.6-volt valves consuming 150 mA and connected them in series across a stabilized h.t. supply. As only 0.3-ampere valves were available this method could not be used in the present case. Barretters or bridge circuits of lamps or of semi-conductor elements were considered, but both performance and efficiency were poor. Stabilizing transformers were unattractive because of the considerable stray field.

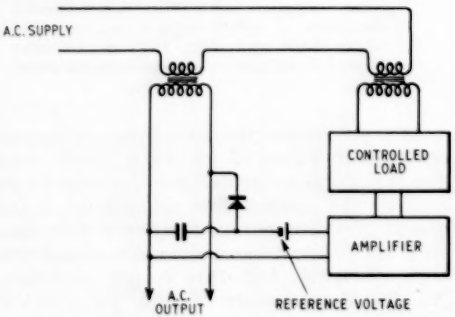


Fig. 33. *Schematic arrangement of a.c. stabilizer of the series-transformer type.*

Vance[50] and Hansen[49] have used series-transformer electronic stabilizers, and a circuit of this type was adopted. The basic arrangement is shown in Fig. 33, and performance in Fig. 34. The stabilization ratio is 20 and the output impedance 6 ohms for a load of 38 volts, 0.9 ampere formed by valve heaters connected in series-parallel as shown in the complete circuit of Fig. 31. The advantage of this type of stabilizer is that the control circuit can operate at an impedance suitable to the load valves available, and the output may be at any desired higher or lower impedance.

The amplification of the stabilizing circuit can be increased by applying positive feedback as shown by the dotted connection, although the advantage is doubtful since the overall stability is reduced. The control voltage is taken from the centre of the smoothing filter and not from the output: this lessens

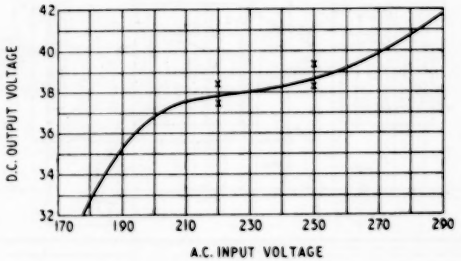


Fig. 34. *Performance of the series-transformer stabilizer with rectifier load and d.c. output. The desired working range is indicated (210-250 V).*

a tendency to instability owing to the very long time constant of the filter.

The four-gap stabilizing valve V_{27} is used in the duplex circuit of Fig. 35, which gives double stabilization of the 70-volt output used as a reference voltage. This is sufficiently stable to enable the h.t. supply shown (+ 500 V) to be taken from the unstabilized circuit which serves the output stages of the signal amplifier. Fluctuations in voltage of

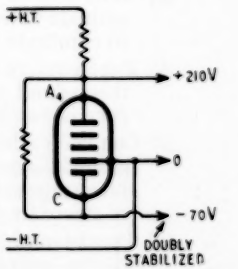


Fig. 35. *Duplex connection of 4-gap, 280-V stabilizing valve to obtain a 210-V stabilized output and a 70-V doubly-stabilized output.*

the stabilizer tube are reduced to a slow rate by an RC filter of 1-second time constant.

Acknowledgment

This experimental design was developed with the assistance of a number of members of the staff of Marconi Instruments Limited and of the Research Laboratories, Marconi's Wireless Telegraph Co., Ltd. Thanks are due to W. Grey Walter, M.A., and Dr. G. D. Dawson for their help in interpreting the requirements of the electro-encephalographer.

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TEMPERATURE-DEPENDENT RESISTORS*

Use as Electric Circuit Elements

By W. H. B. Cooper, A.M.I.E.E. and R. A. Seymour

SUMMARY.—An approximate analysis is given of the behaviour of a temperature-dependent resistor in an electrical circuit; only relatively small changes of temperature are considered. The results are applied to the case of an oscillator which includes a temperature-dependent resistor as a control element.

1. Introduction

Resistors which exhibit non-linear electrical properties, by virtue of a change in temperature with applied power, are finding many applications in telecommunication work. They are particularly useful in circuits which are designed to exercise control over the amplitude of an alternating waveform without introducing non-linear distortion. Such devices, in practice, commonly take the form of "Glow lamps" and "Thermistors."

An approximate theory of operation of such devices is presented and this is followed by an analysis of an oscillator using a temperature-dependent resistor as an amplitude control element.

The following properties of the device are important:—

(*a*) Heat dissipation; due to radiation or conduction, or both.

(In the case of conduction the propagation time of heat transfer might be important, but can be ignored where energy is lost by radiation.)

(*b*) Change of resistance with temperature. (The temperature coefficient may be either positive or negative.)

(*c*) Thermal capacity.

The manner in which these items vary with material, shape and size is not considered.

2. Static Characteristics

Under steady-state conditions the mechanism of heat dissipation is of no consequence, but an expression which fits the radiation case reasonably well is

$$P = AT^{\alpha} \quad .. \quad .. \quad .. \quad (1)$$

where P is power radiated in watts
and T is temperature in degrees absolute.

The change of resistance with temperature is expressed in the form

$$R = BT^{\rho} \quad .. \quad .. \quad .. \quad (2)$$

Some representative values for α and ρ are given in Table I and whilst ρ and in particular α, vary somewhat with temperature, the expressions are adequate for the present purpose.

If a steady-state voltage V is applied, resulting in a current I amperes then

$$P = IV \text{ watts}$$

TABLE I

TEMP. : T Degrees Absolute	Res. at T / Res. at 293°	ρ (Tungsten)	† α	Measured Values of α			
				A	B	C	D
293	1.0	—	—	—	—	—	—
300	1.03	1.23	—	—	—	—	—
350	1.25	1.23	11.4	7.6	7.6	6.9	7.1
400	1.47	1.23	9.5	5.7	4.4	5.6	6.2
500	1.92	1.23	7.6	3.7	3.3	5.0	6.2
600	2.41	1.23	7.1	3.2	3.3	4.5	5.6
700	2.93	1.23	6.7	3.0	3.3	4.5	5.2
800	3.46	1.23	6.3	—	—	—	—
900	4.0	1.22	5.9	—	—	—	—
1000	4.54	1.20	5.6	—	—	—	—

Sample A.—Switchboard Type of Lamp 6 V.
,, B.—Switchboard Type of Lamp 24 V 50 mA.
Sample C.—Osram "Pygmy" 240 V 15 W.
,, D.—Ordinary Vacuum Lamp 240 V 25 W.

* M.S. accepted by the Editor, Oct. 1946.

† Ref. No. 2.

and $R = \dfrac{V}{I}$ ohms

From (1) and (2)

$$V = KI^n \quad .. \quad .. \quad .. \quad (3)$$

where $n = \dfrac{\alpha + \rho}{\alpha - \rho}$

and $$\frac{\partial V}{\partial I} = nKI^{n-1} \quad .. \quad .. \quad (3a)$$

$$= nR$$

It is notable that a negative differential resistance can be obtained even if "ρ" is positive.

It is also interesting to note from equation (3) that when $n = -1$, $VI =$ constant. This is clearly a limiting condition for negative values of n since when $n = -1$ a change of volts or current produces no change of power. It is brought about by making $|\rho| \gg \alpha$, the sign of ρ being unimportant.

The constants can be readily determined experimentally, first of all by using an oven for reasonably low temperatures, and determining ρ with the aid of equation (2) from the experimental results. Similarly T can be determined by means of equation (1) from the voltage/current characteristics.

3. Dynamic Characteristics

Only small changes of temperature are considered. Assume the temperature to change from some value T_0 to $T_0\,(1 + y)$, where y is very small compared with unity, then equations (1) and (2) may be written

$$P = AT_0^{\alpha}\,(1 + \alpha y)$$

$$= P_0\,(1 + \alpha y) \quad .. \quad .. \quad (4)$$

and $$R = BT_0^{\rho}\,(1 + \rho y)$$

$$= R_0\,(1 + \rho y) \quad .. \quad .. \quad (5)$$

It is clear that if the power applied to the device is suddenly changed then the temperature will change only after a time lag since energy is absorbed or released by the thermal capacity H_0. If W watts, the applied power, is changing with time then the energy Wdt supplied in time dt is used up partly in radiation Pdt and partly in changing the temperature by an amount dT.

Hence $Wdt = Pdt + H_0 dT \quad .. \quad .. \quad (6)$

is the basic energy equation assuming energy is lost by radiation and/or conduction, provided the propagation time can be ignored. As small fractional changes "y" are considered

$$T = T_0\,(1 + y)$$

$$\frac{dT}{dt} = T_0 \frac{dy}{dt}$$

If this small change in temperature is brought about by a small change in applied power from W_0 to $W_0\,(1 + \Delta)$ watts

then $$\frac{H_0 T_0}{W_0} \cdot \frac{dy}{dt} = (1 + \Delta) - (1 + \alpha y)$$

$$.. \quad .. \quad (6a)$$

Since $W_0 = P_0$ (a steady-state condition)

$$\therefore \frac{H_0 T_0}{W_0} \frac{dy}{dt} + \alpha y = \Delta \quad .. \quad .. \quad (7)$$

Hence $$y = A \exp\left(-\frac{\alpha W_0}{H_0 T_0} t\right) + \frac{\Delta}{\alpha} \quad .. \quad (7a)$$

It is convenient to substitute

$$\frac{H_0 T_0}{W_0} \equiv L_0 \quad .. \quad .. \quad .. \quad (8)$$

so that $\dfrac{L_0}{\alpha}$ becomes the thermal time constant of the device.

4. Formulae for the Modulation Resistance

A voltage of the form $E_0\,(1 + m \sin \omega t)$, where m is small compared with unity, is applied to a thermally-operated resistor whose resistance varies according to the law $R = R_0\,(1 + \rho y)$ where y is a function of time to be determined. The power supplied will be

$$W = \frac{E_0^2\,(1 + m \sin \omega t)^2}{R_0\,(1 + \rho y)}$$

$$= W_0 \frac{\left(1 + \dfrac{m^2}{2} + 2m \sin \omega t - \dfrac{m^2}{2} \cos 2\omega t\right)}{(1 + \rho y)}$$

where

$$W_0 = \frac{E_0^2}{R_0}$$

Since y is small compared with unity this becomes

$$W = W_0\left(1 + \frac{m^2}{2} + 2m \sin \omega t - \frac{m^2}{2} \cos 2\omega t\right)(1 - \rho y) \quad .. \quad (9)$$

From equations (6), (6a) and (9) it follows that

$$\frac{T_0 dy}{dt} = \frac{W_0}{H_0}\left\{\left(1 + \frac{m^2}{2} + 2m \sin \omega t - \frac{m^2}{2} \cos 2\omega t\right)(1 - \rho y) - (1 + \alpha y)\right\} \quad (10)$$

which, if $\left(\dfrac{m^2}{2} + 2m \sin \omega t - \dfrac{m^2}{2} \cos 2\omega t\right)$ is small compared with unity becomes

$$L_0\frac{dy}{dt} + (\rho + \alpha)y = \left(\frac{m^2}{2} + 2m \sin \omega t - \frac{m^2}{2}\cos 2\omega t\right) \quad .. \quad .. \quad .. \quad (11)$$

This is a linear differential equation and its solution is

$$y = A \exp\left(-\frac{\rho + \alpha}{L_0}\right)t + \frac{m^2}{2(\rho + \alpha)} + \frac{2m \sin \omega t}{(\rho + \alpha) + j\omega L_0}$$

if the harmonic terms are neglected.

Only the steady state solution is required and this is :—

$$y = \frac{m^2}{2(\rho + \alpha)} + \frac{2m \sin \omega t}{(\rho + \alpha) + j\omega L_0}. \quad .. \quad (12)$$

The fractional change of resistance is ρy.

In the case where the device has a resistance added in series, of value equal to R_0 the solution for the modulation of resistance in the steady state for a voltage of the form $E_0(1 + m \sin \omega t)$ applied to the combination is :—

$$\rho y = \frac{\rho m^2}{2\alpha} + \frac{2m\rho \sin \omega t}{(\alpha + j\omega L_0)} \quad .. \quad .. \quad (13)$$

Where the device has in series with it a resistance considerably greater than the resistance R_0, and a current of the form $I_0(1 + m \sin \omega t)$ is passed through the combination, the fractional change of resistance will be

$$\rho y = \frac{\rho m^2}{2(\alpha - \rho)} + \frac{2m\rho \sin \omega t}{(\alpha - \rho) + j\omega L_0} \quad .. \quad (14)$$

5. Measurement of Modulation Resistance

The method of test employed a 300-c/s carrier supply instead of direct current, in order to ease the amplification problems. The circuit layout is given in Fig. 1. The output impedance of amplifier A was low compared with the lamp circuit and the input impedances of amplifiers B and C not less than 100,000 ohms. The 300-c/s carrier could be modulated sinusoidally to a depth of 20% at frequencies from 0.1 c/s upward.

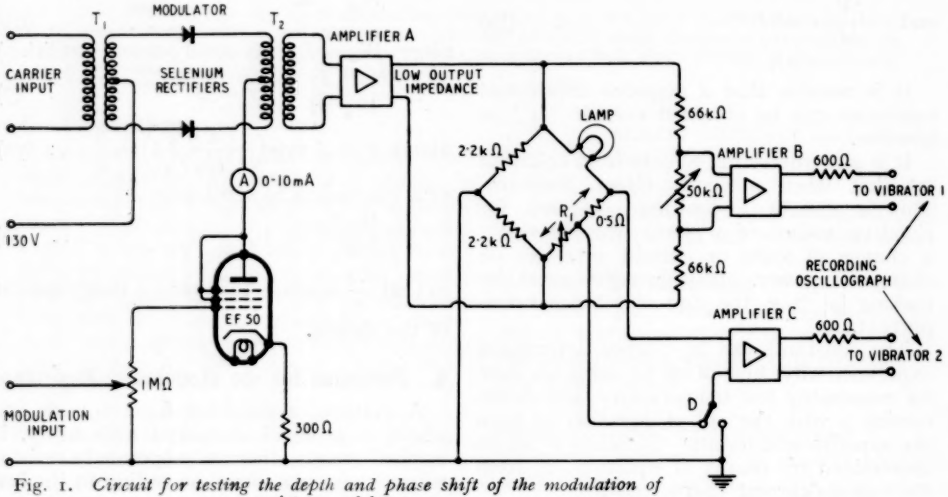


Fig. 1. *Circuit for testing the depth and phase shift of the modulation of resistance of lamps.*

Amplifier B gave an output in phase with the input voltage to the bridge, and amplifier C an output proportional either to the voltage developed across the opposite diagonal of the bridge, or to the current passing through the lamp filament, according to the position of the link D. The bridge method was used when the phase shift was less than 6 degrees and hence difficult to measure by direct comparison. In this case, resistance R_1 was adjusted until a minimum output from the bridge was secured. The following calculation was employed to evaluate the modulation factor for the filament when using the bridge circuit :

For an input voltage of the form $E(1 + m \sin \omega t) \sin ct$ where c is so large that the filament does not respond, the filament

resistance has been shown to respond to the modulation ω in such a way that

$$R = R_0 (1 + \frac{\rho m^2}{2\alpha} + A \sin \omega t) \quad (15)$$

where A the modulation factor, is complex, $A\rho j\phi$, see equation (13), R_0 is the resistance with no modulation, and harmonics are neglected.

Now the current I through the filament and R_1 in series will be

$$I = \frac{E (1 + m \sin \omega t) \sin ct}{R_1 + R_0\left(1 + \frac{\rho m^2}{2a} + A \sin \omega t\right)} \quad (16)$$

When the bridge is adjusted for minimum output $R_1 = R_0\left(1 + \frac{\rho m^2}{2\alpha}\right)$

Hence $I = \frac{E}{2R_0} \cdot \frac{(1 + m \sin \omega t) \sin ct}{[1 + (A/2) \sin \omega t]}$,

neglecting the $\frac{\rho m^2}{2\alpha}$ term, and the voltage across R_1 will equal

$$V = \frac{E}{2} \frac{(1 + m \sin \omega t) \sin ct.}{[1 + (A/2) \sin \omega t]}$$

When $A/2$ is small compared with unity the expression can be simplified and to a first order becomes

$$V = \frac{E}{2}\left(1 + m \sin \omega t - \frac{A}{2} \sin \omega t\right) \sin ct$$

The voltage developed across one of the ratio arms will be

$$V_1 = \frac{E}{2}\left(1 + m \sin \omega t\right) \sin ct$$

The output from the bridge therefore will be $V - V_1$

$$\approx \frac{E}{4} A . \sin \omega t . \sin ct \quad .. \quad .. \quad (17)$$

TABLE II

COMPARISON OF MEASURED AND CALCULATED VALUES OF $|A|$ AND ϕ FOR VARIOUS TYPES OF FILAMENT

Switchboard Type 6 V.

Mean Resistance 24 ohms. Mean Current 6.7 mA.

Frequency of Modulation c/s	m	Measured Values		Calculated Values	
		$\lvert A\rvert$	ϕ	$\lvert A\rvert = \frac{2m\rho}{\sqrt{\alpha^2 + (\omega L_0)^2}}$	$\phi = \lvert \tan^{-1} \frac{\omega L_0}{\alpha}$
0.1	0.176	0.093	$\lvert 17°$	0.127	$\lvert 24°$
0.2	0.16	0.061	$\lvert 45°$	0.096	$\lvert 42°$
0.5	0.16	0.034	$\lvert 72°$	0.053	$\lvert 66°$
1.0	0.16	0.02	$\lvert 79°$	0.028	$\lvert 77°$
1.5	0.16	0.011	$\lvert 82°$	0.019	$\lvert 81° 24'$
2.0	0.16	0.0097	$\lvert 83°$	0.014	$\lvert 83° 30'$

Osram " Pygmy " 15 W.

Mean Resistance 470 ohms. Mean Current 1.9 mA.

0.1	0.143	0.0057	—	0.0076	$\lvert 85°$
0.2	0.143	0.003	$\lvert 86°$	0.0038	$\lvert 86° 48'$
0.33	0.143	0.0022	$\lvert 86°$	0.0023	$\lvert 88°$
0.5	0.143	0.0023	$\lvert 90°$	0.0015	$\lvert 88° 40'$

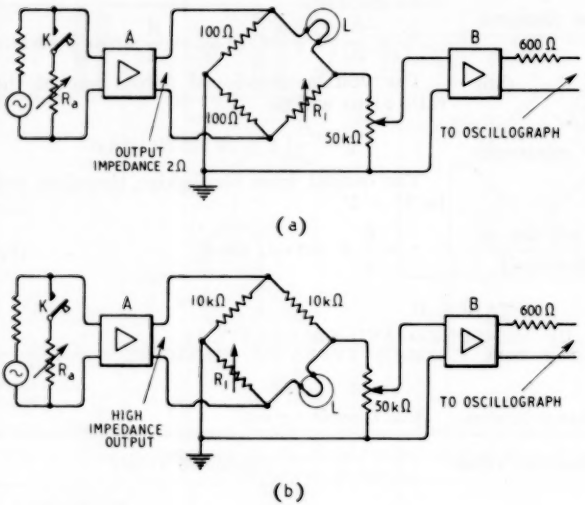
Switchboard Type 24 V 50 mA.

Mean Resistance 86 ohms. Mean Current 5 mA.

0.1	0.107	0.029	$\lvert 64°$	0.048	$\lvert 49°$
0.2	0.107	0.013	$\lvert 73°$	0.03	$\lvert 67°$
0.5	0.107	0.0072	$\lvert 82°$	0.012	$\lvert 80°$
1.0	0.08	0.0056	$\lvert 85°$	0.0048	$\lvert 85°$
2.0	0.08	0.0026	$\lvert 90°$	0.0025	$\lvert 87° 30'$

By calibrating the recording circuit the value of $|A|$ may be determined, and by comparison with the reference trace the angle ϕ can be evaluated.

A comparison of measured and calculated values for $|A|$ and ϕ is given in Table II for the case where the filament has a resistance in series of a value equal to its mean resistance.



6. Measurement of Thermal Time Constant

The methods (*a*) and (*b*) shown in Fig. 2 differ only in that in method (*a*) the lamp filament is supplied from a generator whose resistance is equal to R_0, the resistance of the filament, and in method (*b*) the lamp filament is fed from a high-impedance generator.

In both methods the operating procedure is as follows. The bridge is balanced by the adjustment of R_1 for a particular value of voltage in method (*a*), and of current in method (*b*). With key K first open and then closed, R_a is adjusted to give a certain percentage change in voltage or current. With the key closed the bridge is balanced; then the key is opened whereupon the output from amplifier B rises from zero to some steady value. The envelope of the trace as recorded on the oscillograph* will be of exponential form. The value of this is obtained from the trace as follows:

Method (*a*)—Let the initial input voltage E to the bridge become $E(1+m)$ after opening the key. Let $R_1 = R_0$ the resistance of the lamp filament for a voltage E applied to the bridge and let the filament resistance vary as $R_0(1+\rho y)$ when the voltage is changed to $E(1+m)$, y being a function of time.

The current through the lamp filament will be—

$$\frac{E}{2R_0}(1+m)(1-\tfrac{1}{2}\rho y)$$

and the power W in the lamp filament will be

$$W = \frac{E^2}{4R_0^2}(1+2m)(1-\rho y)R_0(1+\rho y)$$

$$= \frac{E^2}{4R_0}(1+2m) = W_0(1+2m)$$

Hence from equation (6) *et seq.*

$$W_0\{(1+2m)-(1+\alpha y)\}.\,dt = H_0\,dT.$$

and $$\frac{L_0\,dy}{dt} + \alpha y = 2m$$

Fig. 2. *Circuit for the measurement of the thermal time-constant of a filament.*

and when the boundary conditions are such that when $t = 0$, $y = 0$ this becomes

$$y = \frac{2m}{\alpha}\left(1 - e^{-\alpha t/L_0}\right) \quad .. \quad .. \quad (18)$$

The change of resistance by definition is $R_0\rho y$

$$= R_0\frac{2m\rho}{\alpha}\left(1 - e^{-\alpha t/L_0}\right) \quad .. \quad (19)$$

Now the output from amplifier B into the oscillograph will be proportional to the above expression, and thus if the output reaches 0.632 of its final value in N seconds

Then $L_0/\alpha = N$ sec .. .. .. (20)

and since $L_0 = H_0T_0/W_0$

$$H_0 = \alpha W_0 N/T_0 \text{ watt sec/°C} \;..\; (21)$$

Method (*b*)—The solution similarly obtained for the change of filament resistance is

$$R_0\rho y = \frac{R_0 2m\rho}{\alpha-\rho}\left\{1 - \exp\left(-\frac{\alpha-\rho}{L_0}t\right)\right\} \quad (22)$$

and $$H_0 = \frac{W_0(\alpha-\rho)N}{T_0} \text{ watt sec/°C} \quad (23)$$

*A Cambridge Oscillograph was used.

7. Control Element in RC Oscillator

7.1. *Circuit arrangement*

The essential portions of the oscillator circuit are shown in Fig. 3 (*a*). The transmission loss through the selective network

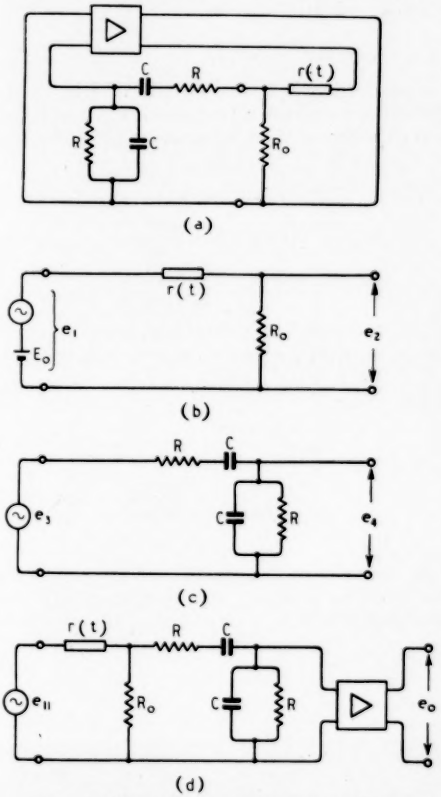


Fig. 3. *Thermally controlled RC oscillator.*

is a minimum and with zero phase shift at that frequency $\frac{c_0}{2\pi}$ c/s at which $c_0 = \frac{1}{CR}$. The drive amplifier is assumed to be perfect from the point of view of freedom from amplitude and attenuation distortion, whilst the lamp circuit elements $r(t)$ and R_0 act as a governor.

Circuit impedances are such that the lamp control circuit is fed from a constant-voltage source, the selective-network impedance at the oscillating frequency is high compared with R_0 and the amplifier input impedance is again high compared with that of the selective network. The voltage loss in the control circuit is approximately two to one when the circuit is oscillating at the design level and that in the selective network shown, three to one, so that the nominal gain of the amplifier is six times. Assume that steady-state oscillating conditions have been reached, then it has been observed that, if the circuit is given an impulse of any sort (e.g., the frequency-change switch operated), the oscillating voltage may be modulated by a low-frequency damped sine wave. An example of this is shown in Fig. 4. Whilst the oscillator circuit has a number of attractive features, particularly for audio-frequency work the duration of the disturbance may be so great as to render the circuit quite unsuitable as a variable frequency test oscillator. The duration of the disturbance is influenced by many factors one of which is the linearity of the drive amplifier. Other things being equal, the perfect amplifier allows the disturbance to last for the greatest length of time.

In the following sections an attempt is made to analyse the circuit using the theory for the temperature dependent resistor as developed in Sections 1 to 4.

7.2. *Thermally operated circuit*

In the circuit of Fig. 3 (*b*) the non-linear element is represented by $r(t)$. A modulated d.c. source represented by e_1, where $e_1 = E_0 (1 + m \sin \omega t)$, is applied to the circuit which has been previously adjusted so that in the absence of modulation (i.e., $m = 0$) the element $r(t) = R_0$

Under these conditions,

$r(t) = R_0 (1 + \rho y)$ where y is the temperature modulation.

From Section 4, Equ. (13) it follows that, under steady state conditions

$$y = \frac{m^2}{2\alpha} + \frac{2m \cdot \sin \omega t}{\alpha + j\omega L_0}$$

where $L_0 = H_0 T_0 / W_0$

and α = radiation index.

The output voltage of Fig. 3 (*b*) will be

$$e_2(t) = \frac{E_0}{2} \frac{(1 + m \sin \omega t)}{\left[1 + \frac{\rho m^2}{4\alpha} + \frac{\rho m}{\alpha + j\omega L_0} \sin \omega t\right]}$$

$$\approx \frac{E_0}{2}\left[1 - \frac{\rho m^2}{4\alpha} + m\left\{\frac{(\alpha - \rho) + j\omega L_0}{\alpha + j\omega L_0}\right\} \sin \omega t\right] \quad (24)$$

In the actual oscillator circuit E_0 will be the r.m.s. value of the oscillation frequency $\frac{c_0}{2\pi}$ c/s, a value so large that the thermal element will not respond cyclically.

7.3 *Frequency-selective RC network.*

A typical form of selective unit is shown in Fig. 3(*c*). This particular network produces a voltage loss $\left(\text{i.e., } \frac{e_3}{e_4}\right)$ of 3 at the frequency of zero phase shift; it is referred to as the 10-db network.

$$c_0 = \frac{1}{CR} \text{ radians/sec.}$$

The steady-state frequency response is given by

$$e_4 = \frac{e_3}{3\left[1 + j\tfrac{1}{3}\left(x - \frac{1}{x}\right)\right]}$$

where $x = \dfrac{\text{Frequency under consideration}}{\text{Frequency of zero phase shift.}}$

The effective Q_0 is only $\frac{1}{3}$.

The steady-state response to a signal of the form $e_3 = E(1 + g \cos \omega t) \cos c_0 t$ is given by :—

$$e_4 \approx \frac{E}{3}\left[\cos c_0 t + \frac{g \cos (c_0 + \omega)t}{2\left(1 + j\frac{2\omega}{3c_0}\right)} + \frac{g \cos (c_0 - \omega)t}{2\left(1 - j\frac{2\omega}{3c_0}\right)}\right]$$

provided

$$c_0 \gg \omega$$

$$\therefore e_4 \approx \frac{E}{3}\left\{1 + g_1 \cos \omega t\right\} \cos c_0 t \qquad (25)$$

where $g_1 = \dfrac{g}{1 + j\omega L_c}$

and $L_c = \dfrac{2}{3c_0}$

The modulation envelope is modified in amplitude and phase by the factor $\dfrac{1}{1 + j\omega L_c}$ in passing through the RC network. Corresponding values of L_c can be found for other types of selective network.

7.4. *Combined response.*

Considering now the circuit of Fig. 3 (*d*) where the amplifier is linear and has a voltage gain of 6 times, then when e_{11} is given by :—

$$e_{11} = E(1 + m \sin \omega t) \sin c_0 t$$

it follows from equations (24) and (25) that

$$e_0 = E \sin c_0 t \left[1 - \frac{\rho m^2}{4\alpha} + \frac{m\{(\alpha - \rho) + j\omega L_0\}}{(\alpha + j\omega L_0)(1 + j\omega L_c)} \sin \omega t\right]$$

For the purpose of the following analysis the oscillation frequency can be regarded as d.c. so that $E \sin c_0 t$ will be replaced by E_0 giving

$$e_0 = E_0\left[1 - \frac{\rho m^2}{4\alpha} + \frac{m\{(\alpha - \rho) + j\omega L_0\}}{(\alpha + j\omega L_0)(1 + j\omega L_c)} \sin \omega t\right] \qquad .. \quad .. \quad (26)$$

If the output terminals are connected to the input as in Fig. 3(*a*), the system will behave like a feedback amplifier with a response

$$\mu = \frac{1}{1 - F(j\omega)} \qquad .. \quad .. \quad (27)$$

where

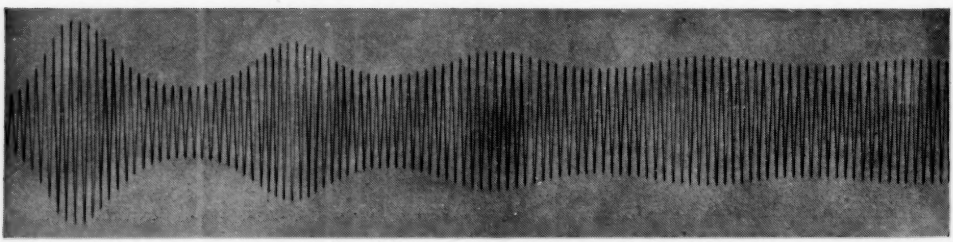


Fig. 4. *Modulation of output of RC oscillator; carrier frequency* 31 *c/s, modulation frequency* 1.26 *c/s.*

$$F(j\omega) = \frac{(\alpha - \rho) + j\omega L_0}{(\alpha + j\omega L_0)(1 + j\omega L_c)} \qquad (28)$$

The system will be stable so long as $F(j\omega)$ does not enclose the point 1, 0 in the complex plane. Equation (27) contains all the information necessary to obtain the response of

the system to a suddenly applied pulse; it is complicated, but with typical values of constants encountered with these circuits some useful approximations can be made. It might be expected that the frequency $\frac{\omega_m}{2\pi}$ at which μ is a maximum would be given by equating the argument of $F(j\omega)$ to zero, when

$$\omega_m^2 = \frac{\rho}{L_c L_0} + \frac{\rho\alpha}{L_0^2} - \frac{\alpha^2}{L_0} \quad .. \quad .. \quad (29)$$

$$\approx \frac{\rho}{L_0 L_c} \text{ for normal circuit values} \quad (30)$$

Equation (30) agrees, to within a few per cent, with experimentally determined values.

From equation (27)

$$\mu = \frac{(\alpha - \omega^2 L_0 L_c) + j\omega (L_0 + \alpha L_c)}{(\rho - \omega^2 L_0 L_c) + j\omega\alpha L_c}$$

From equation (30) and neglecting αL_c compared with L_0

$$\mu \approx \frac{(\alpha - \rho x^2) + j\omega L_0}{\rho(1 - x^2) + j\alpha\omega L_c} \quad \text{where } x = \frac{\omega}{\omega_m}$$

It will be appreciated that the values of μ for values of x around unity are of prime importance so that if

$$(\alpha - \rho) \ll \omega L_0$$

$$\mu \approx \frac{j\omega L_0}{\rho(1 - x^2) + j\alpha\omega L_c}$$

$$\approx \frac{L_0/\alpha L_c}{1 + j/\alpha\sqrt{\frac{L_0\rho}{L_c}}(x - 1/x)} \quad .. \quad (31)$$

Equation (31) is identical in form with that of an LCR circuit in which the Q_0 (Q at resonance) is given by

$$Q_0 = \frac{1}{\alpha}\sqrt{\frac{L_0\rho}{L_c}} \quad .. \quad .. \quad (32)$$

and it follows that the envelope of the response to a suddenly applied pulse will be given by

$$\text{Envelope of response} = (1 - e^{-t/\tau}) \quad (33)$$

where the time-constant

$$\tau = \frac{2Q_0}{\omega_m}$$

$$\approx \frac{2L_0}{\alpha} = \frac{2H_0 T_0}{W_0\alpha} \quad .. \quad .. \quad (34)$$

This is independent of the oscillator frequency $\frac{c_0}{2\pi}$; it is twice the thermal time-constant of the lamp. In view of the tuned circuit analogy this is not surprising.

The decrement is given by

$$\text{Decrement} = \text{antilog } [0.43\pi/Q_0] \quad .. \quad (35)$$

A direct method of measuring the time constant is given in Section 6.

Some calculated and measured values of ω_m and decrement are given in the Table III for an oscillator incorporating a "10-db" RC network and a 6.0-V switchboard-type of lamp operating with an unmodulated current of 5.0 mA r.m.s.

$\alpha = 3.6$

$L_c = 2/(3\,c_0)$

$= 0.0035$ for 30 c/s

$= 0.0021$,, 50 c/s

$= 0.0011$,, 100 c/s

$\rho = 1.2$

$L_0 = \frac{H_0 T_0}{W_0} = 6.2$

TABLE III

$c_0/2\pi$	ω_m		Decrement	
	Measured	Calculated	Measured	Calculated
30 c/s	7.9	7.4	2.1	1.3
50	9.6	9.6	3.5	1.2
100	13.4	13.5	5.2	1.15

Whilst measured resonant frequencies are in good agreement with the theory, the measured time constants are very much shorter than those calculated. An attempt is made, in the next section, to explain this discrepancy.

7.5 *Decrement as a Function of Modulation Depth "m"*

In the expansion which gives equation (24) the term $-\frac{\rho m^3}{4\alpha}\sin\omega t$ has been omitted and if this included equation (28) becomes

$$F(j\omega, m) = \frac{(\alpha - \rho) + j\omega L_0}{(\alpha + j\omega L_0)(1 + j\omega L_c)} - \frac{m^2\rho}{4\alpha(1 + j\omega L_c)} \quad .. \quad .. \quad (36)$$

From this point onwards the analysis again follows on the same lines as that of Section 7.4 but since equation (36) represents a non-linear system the results obtained this way are open to question.

$$\mu \approx \left[\frac{\rho m^2}{4\alpha} + \frac{\alpha L_c}{L_0}\right]\left[1 + \frac{j\sqrt{L_0 L_c \rho}\left\{x - \frac{1}{x}\left(1 + \frac{m^2}{4}\right)\right\}}{L_0\frac{m^2\rho}{\alpha} + \alpha L_c}\right]^{-1} \quad .. \quad .. \quad .. \quad .. \quad (37)$$

where $x = \omega/\omega_m$ and $\omega_m^2 = \rho/L_0L_c$ } as before

This expression of course reduces to (31) when $m = 0$. Considered as a function of x, equation (37) does not exhibit the symmetry of a normal resonance curve, but for values of x greater than unity, where the factor $\left(1 + \frac{m^2}{4}\right)$ becomes less important, the response will be similar to that of a normal resonant circuit with a Q_0 value given by

$$Q_{0m} = \frac{\sqrt{L_0L_c\rho}}{L_0\frac{\rho m^2}{4\alpha} + \alpha L_c} \quad .. \quad .. \quad (38)$$

In which case the time constant will be

$$\tau_m = 2Q_{0m}/\omega_m$$

$$= \frac{2L_0}{\alpha + \frac{L_0\rho m^2}{4L_c\alpha}} \quad .. \quad .. \quad (39)$$

$$= \frac{2L_0}{\alpha\left\{1 + \frac{m^2}{4}Q_0^2\right\}} \quad .. \quad .. \quad (40)$$

and the decrement

$$= \text{antilog}\left[0.43\pi\left(\frac{1}{Q_0} + \frac{m^2Q_0}{4}\right)\right] \quad (41)$$

where $Q_0 = \frac{1}{\alpha}\sqrt{\frac{L_0\rho}{L_c}}$ [equation (32)]

Table IV gives calculated values of decrement, from equation (41) for different depths of modulation.

TABLE IV

$c_0/2\pi$	Q_0	Decrement (calculated)			
		$m = 0$	$m = 0.1$	$m = 0.2$	$m = 0.3$
30 c/s	12.8	1.3	1.4	1.9	3.2
50	16.5	1.2	1.4	2.1	3.9
100	23.5	1.15	1.4	2.5	6.2

The time constant τ_m of the disturbance may also be written

$$\tau_m = \frac{2L_0}{\alpha} \cdot \frac{1}{\left\{1 + \frac{m^2}{4}\left(\frac{L_0}{\alpha}\right)^2\omega_m^2\right\}} \quad (42)$$

where $\omega_m^2 = \frac{\rho}{L_0L_c}$ [equation (30)]

ω_m being the modulation frequency.

Clearly the time constant $\frac{L_0}{\alpha}$ should be as short as possible consistent with the condition that it should be long compared with the period of the oscillation frequency; in addition it is advantageous to make ω_m as large as possible. Whilst Table IV does not agree very accurately with experimental results the trend is in the right direction and the deductions from equation (42) are in accord with practice.

Acknowledgment

The authors are indebted to the Engineer-in-Chief of the British Post Office for permission to publish this article.

REFERENCES

[1] "On the Theory of Light Modulation of AC heated Glow-Lamps in the Audio Frequency Range," by Horst Kohler, *E.N.T.*, April 1938.

[2] "The Characteristics of Tungsten Filaments as Functions of Temperature," by Dr. H. Jones and I. Langmuir, *Gen. elect. Rev.*, Vol. 30, pp. 310, 354, 408.

NEW BOOKS

Electronics

Edited by BERNARD LOVELL. Pp. 660 + xvi with 404 illustrations. Pilot Press, Ltd., 45, Great Russell Street, London, W.C.1. Price 42s.

There are fourteen chapters by thirteen different authors in this book. The chapters are all on different subjects having in common only their dependence on electronics. The book is, therefore, essentially a collection of monographs on related subjects and there will be few individuals to whom the whole will be of equal interest.

The collection of such a diversity of material in a single volume is not without advantages, however, if only because it leads the specialist to read beyond his own narrow subject. It gives him an insight into many dissimilar problems and by doing so helps him to achieve a better perspective of his own.

Chapter I "Electron Physics" is by Dr. F. A. Vick of the University of Manchester. It comprises a review of 51 pages of the present state of knowledge of electron physics, "including the behaviour of electrons in metals and non-metallic solids." Thermionic, field, and photo-electric emission are dealt with as well as contact potentials, the luminescence of solids and photo-conductivity.

Chapter II is by Dr. H. G. Lubszynski of the E.M.I. Research Laboratories and is entitled "Photo-Cells for the Visible and Ultra-Violet." There are 43 pages covering photo-emissive, photo-conductive, and photo-voltaic cells, and gas-filled and electron-multiplier types are included. Several pages are devoted to noise and applications. In spite of the large amount of ground covered quite a lot of detailed information is included.

The discussion of photo-cells is continued for another 37 pages in Chapter III, "Recent Advances in Photo-Cells for the Infra-Red" by Dr. A. Elliott, formerly of the Admiralty Research Laboratory. This deals chiefly with thallium-sulphide, lead-sulphide, lead selenide and telluride photo-conductive cells.

Following this there is an extraordinarily good chapter (IV) on "Electronic Generation of Tele-

vision Signals " by Dr. J. D. McGee of the E.M.I. Research Laboratories. In the first nine of his 77 pages he briefly describes early methods of scanning, such as the Nipkow disc and Campbell-Swinton's early electronic suggestions, and reviews the Farnsworth image dissector and the intermediate-film method. There are then some five pages on the charge-storage principle. The rest of the chapter covers in considerable detail the operating principles and characteristics of the Emitron (iconoscope), Super-Emitron, Orthicon and Super-Orthicon. The author gives a very clear and well-balanced discussion of the characteristics of, and difficulties inherent in, each.

Chapter V is " Thermionic Valves for Very High Frequencies " by Dr. F. C. Thompson of Marconi Valve Laboratories ; 25 pages are devoted to klystrons and magnetrons. The material is largely descriptive and gives a good general picture of the way in which such valves operate.

Chapter VI is of 81 pages ; it is " Radar " by Dr. R. A. Smith of Telecommunications Research Establishment. The fundamentals of propagation and echoes are covered, as well as the frequency spectrum of pulses and the effect of bandwidth on pulse shape. Frequency- and pulse-modulation systems are compared and anomalous propagation (super-refraction) is dealt with. Display systems, responders and aerial scanners are covered. The chapter gives a very good technical picture of radar, but there is little detailed information on circuits.

Chapter VII by L. Atkinson of Ferranti, Ltd., is of 28 pages. " Control Applications of Cold Cathode Valves " covers ignitions, servotrons, stroboscopic lamps, pulse-trigger valves, and the megalite lamp. Circuits of control apparatus are given and include counters, magnetizing equipment, welders, stroboscopes, depth sounders and apparatus for high-speed photography.

" High-Frequency Heating " is dealt with in Chapter VIII by H. Wood of Ferranti, Ltd., in 33 pages. The theories of induction and dielectric heating are given briefly and there are circuits and details of the apparatus employed.

In Chapter IX the same author deals with " A Moisture Content Control Equipment." This is of only 7 pages but includes the principle of operation and the circuit of the apparatus.

" Electronics Applied to Servo-Mechanisms " is the title of Chapter X of 52 pages by F. H. Belsey of Metropolitan-Vickers Electrical Co. It is an extremely good account of the theory of servo-mechanisms and particular stress is laid upon the conditions which are necessary for obtaining stability. The analysis is mathematical.

Dr. L. G. Grimmett, formerly of the Radio-therapeutic Unit of the Medical Research Council contributes Chapter XI " Electronics in Medicine." It is of 38 pages and covers high-energy radiations, such as those produced by X-ray generators, electrostatic generators, cyclotrons, betatrons, synchrotrons, microtrons, linear resonance accelerations, and waveguide accelerators. Diathermy, electron optics, and electro-encephalography are also covered. The chapter is in the nature of a general survey of the subject.

" Electronics in Physiology " is the title of Chapter XII by Dr. R. J. Pumphrey of the Department of Zoology, Cambridge. In 19 pages he describes the use of electronics in the study of nerves and muscles.

Chapter XIII, " The Betatron " is by Dr. J. D. Craggs, of the Metropolitan-Vickers Electrical Co. Of 31 pages it includes the theory of the betatron and describes its construction ; its applications to deep therapy, to nuclear physics and to radiography are also dealt with.

Chapter XIV of 107 pages is the longest in the book. By Dr. V. E. Casslett of the Electrical Laboratory, Oxford, it is entitled " Electron Microscopy and Electron Diffraction."

It includes a good deal of information about electron optics as applied to the electron microscope and the construction of the electron microscope is described. Stabilizing and supply circuits are covered.

Each chapter includes a bibliography and there is a good subject index. The book is well produced and covers an enormous range of materal. Since each chapter is written by a specialist there are remarkably few errors. Its title is perhaps apt to be misleading, for the greatly overworked word " electronics," although used here quite correctly, is likely to convey something different to many.

W.T.C.

Klystron Tubes

By A. E. HARRISON. Pp. 271 + x. McGraw-Hill Publishing Co., Ltd., Aldwych House, London, W.C.2. Price 17s. 6d.

This book, as mentioned in the preface, grew out of the Klystron Technical Manual, which was widely distributed during the war by the Sperry Gyroscope Company to users of klystrons in the United States and also over here in Great Britain. It has been completely revised and much new material has been added, a good deal of which no doubt is based on the author's original contributions to our knowledge of the klystron tube.

No prior knowledge of the klystron tube is assumed and the author is to be congratulated on his extremely clear and readable introduction to a technique,—if not science,—which breaks away so radically from all radio techniques in use right up to the outbreak of war. Once the three or four essentially new principles, which form the basis of all electron tubes conceived in the last ten years, have been grasped,—and these principles are by no means more difficult to understand than those on which, say, triodes are based, most of the microwave-tube technique as applied in the war is intelligible to anybody who has only an elementary knowledge of radio and physics. These principles, in the order in which they emerged in the literature, are : the principle of an inductive coupling between an a.c. component in an electron beam and an external circuit ; the principle of the formation of such an a.c. component, or of " bunching ", by means of the drifting of electrons which have been given an a.c. component of velocity ; and the principle of producing regions of very high r.f. field strengths in circuits by making them all-enclosed ; i.e., cavities. These principles had been in the air for some time, and devices making use of one or two of them had been described. But it was not until the famous series of papers on the klystron by the Varian Brothers, W. W. Hansen and D. L. Webster was published, all in 1939, that it became obvious that a real and great step forward in radio technique had been made. It is no exaggeration to say that the whole of the centimetre techniques, which eventually culminated

in centimetre radar owe a good deal to the klystron in one way or another. Even the development of the magnetron, as invented by Boot and Randall in Birmingham, made a great and significant stride when Sayers applied a measuring technique to it which involved the use of a klystron.

At present, klystrons are virtually supreme in any laboratory which uses microwave techniques. They form the heart of great projects for microwave-transmission systems. They are absolutely essential in any navigation or blind-landing system at microwave frequencies. They form part of all kinds of radar systems. Therefore, a book as authoritative, comprehensive and at the same time as compact, as the one under review must be very welcome to many users and potential users, as well as to designers and investigators, of klystrons all over the world.

The headings of the chapters give a good idea of the ground covered in the treatise: 1. Klystron Construction. 2. Cavity resonators. 3. Electron-bunching theory. 4. Klystron amplifiers. 5. Klystron frequency multipliers. 6. Reflection bunching. 7. Reflex oscillators. 8. Two-resonator oscillators. 9. Multiple resonator tubes, with the sub-headings: oscillator-buffer tubes, cascade-amplifier klystrons, amplifier-multiplier tubes. 10. Modulation of Klystrons, with the sub-headings: frequency modulation, phase modulation, amplitude modulation, pulse modulation, klystron detectors. 11. Klystron tuners. 12. Klystron operation. 13. Klystron power supplies. 14. Microwave measurement techniques, with the sub-headings: Impedance measurement, power measurement, frequency and wavelength measurements, some typical microwave experiments.

There are also a number of valuable klystron design charts, particularly useful to designers of klystrons, and a fairly comprehensive bibliography. Very few references are made in the text, and a minor criticism can be made, namely, that work at places other than at Stanford University and the Sperry Gyroscope Company receives only scant mention.

Though much of the mathematical analysis of bunching, and the theory of the klystron as amplifier, oscillator and frequency multiplier is given in great detail, it is a little unfortunate that the problem of the coupling-out of r.f. power from the bunched-electron beam has not been treated more rigidly, and perhaps a little more convincingly. Also a quantitative treatment of beam loading and space-charge de-bunching would help in enabling a reader new to the subject to form at least an estimate of the order of magnitude of the effects. Designers of klystrons might expect information on the problem of the maximum obtainable beam current and the related electron-optical problems in a book of this kind, perhaps at the cost of the chapter on microwave measurement techniques, which, in my opinion, attempts too much in too short a space. On the other hand, a description of measurements of parameters peculiar to the klystron, such as of the shunt resistance of resonators or of the noise factor of the tube as an amplifier would be apt and welcome in a treatise on the klystron.

It is to be hoped that this book will go through many editions, and that in them the author will deal with the points referred to above.

R.K.

Kettenförmige Ultrakurzwellen-Bandfilter aus Quasistationären Schwingtöpfen

By FRIDOLIN STAUB. Pp. 89 with 43 Figs. Verlag AG. Gebr. Leemann & Co., Zurich. Price 10 francs (Swiss).

This is a doctorate thesis of the Zurich Technischen Hochschule. It deals with the use of cavity resonators in the construction of ultra-short wave band-pass filters. The first 50 pages deal with theory, the remainder with the design, construction and tests of various types of such filters. The filter theory is based on the work of Cauer. Very ingenious arrangements are described in which a number of coupled-cavity resonators are comprised in a single metal cylinder. The frequency employed is about 175 Mc/s and the dimensions of the cavity resonators are such that the conditions may be assumed to be quasi-stationary. The resonators are so enclosed that they do not radiate and they contain no dielectric; they are made of brass or gun-metal, but either coppered or silvered and highly polished on the current-carrying surface. In this way the losses can be kept down to a very low value with the result that very high-class filters can be constructed. The methods of measuring the Q values, bandwidths, etc., are described. This is certainly a valuable addition to the literature of the subject. G. W. O. H.

Radio-Technik, Theorie und Praxis

By Dr. J. DÜRRWANG. Pp. 216 with 168 Figs. B. Wepf & Co., Basel. Price 12 francs (Swiss).

This is the fifth edition; it has been improved and brought up to date by the addition of a chapter on ultra-short waves. It is a non-mathematical book intended for seriously minded amateurs and and radio-technicians. It is well written and well illustrated. The eighteen chapters cover the whole field of radio from Ohm's law to Radar, including Television, every item being explained very clearly in an elementary way with the aid of models and diagrams. A pleasing feature is the meticulous care that has been taken with the nomenclature, and symbols, although in some of the diagrams mA becomes Ma. G. W. O. H.

Introducing Radio Receiver Servicing

By E. M. SQUIRE. 2nd Edition. Pp. 144 + vii. Sir Isaac Pitman & Sons, Ltd., Parker Street, Kingsway, London, W.C.2. Price 7s. 6d.

Welding of Plastics

By G. HAIM and H. P. ZADE. Pp. 206 + xv with 58 illustrations. Crosby Lockwood & Son, Ltd., 20, Tudor St., London, E.C.4. Price 21s.

Mainly devoted to the characteristics of plastics and general methods of welding them. Two of the seventeen chapters deal with high-frequency methods.

GRIFFIN & TATLOCK

A new 600-page catalogue, No. 52 AB, of scientific apparatus, has been produced by Griffin & Tatlock of Kemble Street, Kingsway, London, W.C.2. It is intended primarily for use in educational laboratories and it covers apparatus for metrology, mechanics, geophysics, heat, light, sound, magnetism, electricity including atomic physics, microchemistry, physical chemistry and biology.

CORRESPONDENCE

Letters of technical interest are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Stabilizing Direct-Voltage Supplies

To the Editor, "Wireless Engineer"

SIR,—Dr. Hughes' article in your August issue will be of little value in my opinion until the voltage of gas-filled regulating valves is both reproducible and permanent to within, say, 0.5 volt or less.

Considering regulating valves commercially available, voltage variations of ± 10 per cent exist between samples of any given type, and the voltage of any valve selected at random may vary between 1 per cent and 10 per cent over the permissible valve current range. The voltage will vary between −10 per cent and + 20 per cent during 1,000 hours use. Further erratic jumps in voltage occur, and also drifts as valves heat and cool.

The S130 valve, used in Dr. Hughes' tests, is typical of other available valves; its voltage varies between 115 and 135 for new samples, and by between 0 and 10 volts over the normal current range. Thus two valves not specially selected, are not likely to operate from the 250-volt supply used in the tests. The close agreement between his calculated and observed values is because the "calculated" values are based on "observed" characteristics of the two valves used, so that the results only apply to these two valves, and will be in error if the valves are operated for some time.

The simplest solution to this problem is the most useful. The series-resistor value should be obtained from the following information:—(1) minimum supply voltage, (2) maximum valve voltage, (3) maximum load current, and (4) minimum valve current. Then the maximum valve current must be checked under the other extreme condition:—(1) maximum supply voltage, (2) minimum valve voltage, and (3) minimum load current. If the valve is not over-run, the output voltage and the variation of voltage with load are determined by the specified values for the valve.

The method of obtaining the necessary high striking voltage in the presence of load, as described in Section 4 of the article, is ineffective if, while the regulator is working, the load increases momentarily and extinguishes the glow, or if the supply voltage increases relatively slowly when switched on, as in the case of most valve rectifier supplies.

In July 1941 I approached the manufacturer to have a second anode introduced into the S130. The new valve, the CV45, has characteristics similar to those of the S130, though the connection of the auxiliary anode, through a resistor, to at least 190 volts above cathode, ionizes the gas and allows the main anode to strike at the actual running voltage. This eliminates the striking voltage problem as there is generally more than 190 volts across the reservoir capacitor in the power supply. Similar striking anodes have been fitted in valves since developed. Where two or more valves are connected in series, no higher voltage is required as they can be connected so that twice the running voltage is available in each valve, to cause the auxiliary anode to establish the ionizing glow.

In conclusion, it is to be hoped, now that the causes of the variations in voltage with use, and between samples, are known[1] that British manufacturers will improve their technique and will produce gas filled valves which really are "voltage regulating valves."

WM. H. P. LESLIE.

N. Farnborough, Hants.

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[1] *Philips Research Reports*, Vol. 1.

F. M. Penning and J. H. A. Moubis. "Contraction phenomenon in a neon glow discharge with molybdenum cathode." No. 2, p. 119.

T. Jurriaanse, F. M. Penning, and J. H. A. Moubis. "Normal cathode fall for molybdenum and zirconium in the rare gases." No. 3, p. 225.

T. Jurriaanse. "Influence of gas density and temperature on the normal cathode fall." No. 6, p. 407.

Partially-Screened Open Aerials

To the Editor, "Wireless Engineer"

SIR,—In reply to Señor Colino's letter in the August issue, I wish to state that his picture of the system is consistent with my analysis[1]. Some more detailed discussion of the point raised was given in my paper on the screened-loop[2] where the "inner" and "outer" modes of propagation in a screened system were discussed.

Considering the partially-screened open aerial, we note that equations (1) give for the "inner" mode:

$$\frac{\partial}{\partial x}(v_1 - v_2) = -j\omega(L_{11} - L_{22})i_1$$

and

$$\frac{\partial i_1}{\partial x} = -j\omega C_{11}(v_1 - v_2)$$

showing that the potential difference $(v_1 - v_2)$ between the conductors and the current i_1 flowing up the inner conductor and down the inner surface of the screen are exactly as in the coaxial transmission line which they form; this mode is independent of i_2 and of the medium outside the screen.

Similarly we have

$$\frac{\partial}{\partial x}(i_1 + i_2) = -j\omega(C_{22} - C_{11})v_2$$

and

$$\frac{\partial v_2}{\partial x} = -j\omega L_{22}(i_1 + i_2)$$

corresponding to the "outer" anode. The potential v_2 of the screen and the current $(i_1 + i_2)$ flowing in the outer surface of the screen are propagated in a manner independently of the conditions within the screen.

Thus the screen supports a current $(-i_1)$ flowing on its inner surface and $(i_1 + i_2)$ flowing on its outer surface, the resultant i_2 being the quantity which is significant in reception or transmission. If the inner and outer radii of the screen are r_1 and r_0 the effective circulating current of i_1 in the screen neutralizes that portion of the magnetic field of the inner between r_1 and r_0 and so reduces L_{11} by $2 \log (r_0/r_1)$ e.m.u. per cm.

The fact that any break in the screen contributes

to the pick-up of a wave on the inner conductor has been emphasized in both papers[1,2].

National Physical Laboratory. R. E. BURGESS.
Teddington, Middx. Radio Division.

[1] R. E. Burgess: "Partially-Screened Open Aerials," *Wireless Engineer*, May 1947, Vol. 24, pp. 145-149.

[2] R. E. Burgess: "Screened Loop Aerials," *Wireless Engineer*, May 1944, Vol. 24, pp. 210-221.

Single-Phase and Polyphase Filtering Devices Using Modulation

To the Editor, "Wireless Engineer"

SIR,—In your May issue,* Mr. N. F. Barber describes a method of filtering by modulation, solving in a very elegant manner a problem which could hardly be solved in the conventional way. It may be of some interest to note that the proposed arrangement can be regarded, as a whole, as a set of polyphase modulating and filtering devices, while the simpler system referred to in the introduction of the paper as "the simplest process of filtering by modulation," appears as the corresponding single-phase arrangement. The correspondence would be complete if in the case of this simpler process the output of the filter were translated again into the region of the higher frequencies, by means of a second modulation.

It seems to be worth while discussing, from a general point of view, why the more complicated polyphase system is required for solving the proposed problem, and upon this question I cannot fully agree with Mr. Barber's statements. In his paper, the disadvantages arising from using the simpler system are quoted as follows:

(a) the phase angle of the input voltages is lost;

(b) voltages having the same frequency as the first modulating voltage can be lost.

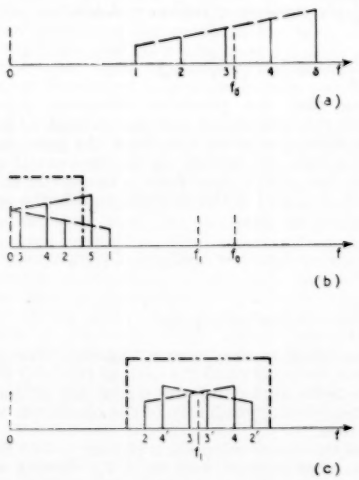

Fig. 1.

The disadvantage quoted under (a) is not connected with the number of phases of the system, but rather with the fact that, by the simpler method, the output voltage has not the same frequency as the corresponding input voltage. Hence this limitation has no interest for the present question. One would conclude, therefore, that the only disadvantage arising from the use of a single-phase set would be the one quoted under (b).

I believe it is worth while emphasizing that a more fundamental disadvantage must be taken into account; namely, the fact that, by the single-phase arrangement, the spectrum of the output voltage would be *unavoidably* accompanied by another disturbing spectrum. This can be explained very simply if one regards a polyphase modulator as an algebraic frequency adder. The frequency of

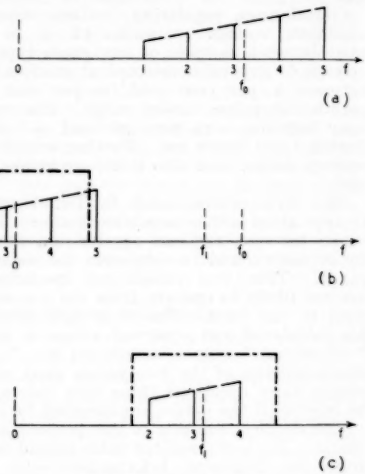

Fig. 2.

the polyphase voltages involved, either input or output, can be regarded as an algebraic quantity and is consequently defined both in value and sign (the sign being related to the sequence). However, the frequency of single-phase voltages can only be defined in value, not in sign.

From this standpoint, the problem involved in Mr. Barber's method can be discussed as follows: a frequency spectrum is translated from the neighbourhood of a frequency f_0 to the neighbourhood of zero frequency. It is here filtered, and then translated again to the neighbourhood of the frequency f_1 (which can eventually be equal to f_0). If this is accomplished by conventional single-phase modulators, then in addition to the disadvantages pointed out by Mr. Barber, the lines of the original spectrum [Fig. 1(a)], which lie on opposite sides of the first modulating frequency f_0, are treated by the first modulation exactly in the same manner [Fig. 1(b)], and then, by the second modulation, added to and subtracted from the second modulating frequency f_1 [Fig. 1(c)]. The output spectrum is, therefore, accompanied by its mirror image with respect to f_1.

Now suppose that one of the inputs of the first modulator is polyphase, and that the same holds for its output. In this case, the lines of the original spectrum lying on opposite sides of the first modulating frequency f_0, [Fig. 2(a)] give, by the first modulation, polyphase voltages of opposite sequence, which can be regarded as voltages of opposite

* Narrow Band-Pass Filter Using Modulation, Vol. 24, p. 132.

frequency, and therefore represented by lines which lie on opposite sides of the zero frequency [Fig. 2(b)].

Suppose furthermore that both inputs of the second modulator are polyphase. Then the frequencies of the spectrum Fig. 2(b) are added to, or subtracted from, the second modulating frequency f_1, according to their sign (i.e., to the sequence of the corresponding voltages) and so the lines of the final spectrum receive their proper allocation [Fig. 2(c)].

It can be concluded, therefore, that the necessity for employing the more elaborate polyphase system is by no means a casual one, but is strictly connected with the general features of polyphase and single-phase systems. Thus Mr. Barber's paper appears to be an excellent example of a comparison of the general performances to be expected from these two types of arrangements. It is obvious that the same results could be predicted, in any particular case, without introducing the conception of polyphase modulating systems, by merely analysing the assembly of single-phase modulators. The analytical treatment is, moreover, of invaluable help in studying the details of the set, the causes of errors and so on. I chose, however, the more synthetic treatment using the conception of polyphase systems, because of its power in predicting, in accordance with general rules, what can be done with a certain device, and what cannot.

G. B. MADELLA.

Instituto Elettrotecnico Nazionale Galileo Ferraris Torino, Italy.

Degrees for Ex-Servicemen

To the Editor, "Wireless Engineer"

SIR,—With regard to my previous correspondence in connection with Degree Courses for ex-servicemen and those engaged in industry, the results of the enquiry have been sent to the Ministry of Education, who have handled the matter sympathetically. They are communicating with the local Councils concerned with a view to providing classes at times more convenient for those members of the staff of firms who cannot be released for day-time attendance, except on Saturday mornings. It must be realized, however, that the final responsibility for the organization of courses rests with the individual teaching establishments.

It is hoped that suitable arrangements can be made in time for the forthcoming winter sessions but it is regretted that it is only in certain parts of the country that the numbers of applications are sufficient to permit any action to be taken.

I would like to thank those who have helped by filling in my questionnaires and to apologise for being able to assist only a portion of them to obtain suitable courses.

O. S. PUCKLE,

Langley Park, Nr. Slough, Bucks. R.F. Equipment, Ltd.

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 1/- each.

ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

582 942.—L.F. selective relay in which the control signals are applied to two cold-cathode tubes, one of which responds to the beginning, and the other to the end of the signal-train.

Standard Telephones and Cables Ltd., F. H. Bray and L. R. Brown. Application date 11th February, 1944.

583 808.—Remote-control switching system for selecting one of several broadcast items distributed over a telephone network.

G. F. Wittgenstein. Application date 13th July, 1943.

584 354.—Thermo-plastic spider for centring the moving coils of a loudspeaker.

H. C. Willson. Application date 8th December, 1945.

AERIAL AND AERIAL SYSTEMS

583 496.—Aerial comprising resonant transmission-line elements the overall length of which is only one-fifth the operating wavelength, particularly for use on aircraft.

Standard Telephones and Cables Ltd. and E. O Willoughby. Application date 21st April, 1944.

584 900.—Flexible streamlined aerial, comprising two wires or strips, for the roof of a motor-car.

E. St.J. Chesney. Application date 5th February, 1945.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

583 758.—Mechanical switching arrangement for reversing the aerial and indicator connections in an automatic direction-finder.

F. Tench (communicated by Automatic Electric Laboratories Inc.). Application date 30th November 1944.

583 787.—Blind-landing system in which an extended aerial system for defining the glide-path is combined with radiolocation equipment for indicating the point-to-point elevation of the aeroplane.

Marconi's W.T. Co. Ltd. and S. W. H. W. Falloon. Application date 5th October, 1939.

583 792.—Logarithmically-graded potentiometer for centring the echo-signal on the time-base of a radiolocation set, and so giving a direct indication of the range of the target under observation.

J. D. Cockcroft and J. W. Pletts. Application date 5th June, 1942.

583 996.—Stabilizing a multivibrator as used for gating or selecting a particular pulse-train from other trains, as in radiolocation equipment.

Standard Telephones and Cables Ltd. and C. W. Earp. Application date 15th September, 1944.

584 046.—Modulation circuit for navigational beacons of the kind in which one of the aerials is energized by sideband energy only.

Standard Telephones and Cables Ltd. (assignees of A. Alford and G. K. Paterson). Convention date (U.S.A.) 15th November, 1943.

RECEIVING CIRCUITS AND APPARATUS

584 082.—Cathode-follower circuit used as a phase-shifter, particularly for feeding signals of opposite polarity to the deflecting plates of a c.r. indicating tube.

Standard Telephones and Cables Ltd., (assignees of H. E. Beste). Convention date (U.S.A.) 6th December, 1943.

584 181.—Circuit design and operating conditions for increasing the overall gain of a transducer or power-amplifier using valves of the tetrode type.

B. M. Hadfield. Application date 15th December, 1944

584 191.—Phase-inverting bridge-circuit, particularly for coupling an unbalanced amplifier to a balanced push-pull stage.

Marconi's W. T. Co. Ltd., (assignees of H. W. Berry). Convention date (U.S.A.) 27th July, 1943.

584 218.—Automatic gain-control system in which regulation is effected by varying the heating current, and therefore the cathode-emission and effective impedance, of a valve.

Siemens Bros. & Co. Ltd., and J. H. Mole. Application date 15th December, 1944.

584 599.—Releasable stop-device for a slow-motion tuning-control to facilitate readjustment to a selected position, and to offset backlash.

H. S. Hall and Pye Ltd. Application date 9th January, 1945.

TRANSMITTING CIRCUITS AND APPARATUS

584 249.—Device for measuring the field-intensity inside a waveguide in terms of the angular displacement of a suspended disc.

E. C. Cork. Application date 25th September, 1941.

584 444.—H.F. impedance-matching transformer in which quarter-wave transmission-line elements are used to offset changes in the mean operating frequency.

J. Collard. Application date 3rd June, 1944.

584 465.—Frequency-modulating system in which the resonance of a crystal-oscillator is controlled by a variable-reactance valve.

Western Electric Co. Inc. Convention date (U.S.A.) 16th September, 1943.

584 642.—Radio-telegraphy transmitter in which the two valves of a multivibrator circuit are alternately blocked and unblocked to generate square-shaped signalling pulses.

Standard Telephones and Cables Ltd., M. M. Levy and E. A. Rattue. Application date 15th January, 1945.

584 942.—Arrangement of a discharge-tube for switching or attenuating the energy flowing in a waveguide.

G. E. F. Fertel and C. S. Wright. Application date 7th June, 1944.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

584 427.—Pulsed carrier-wave system with synchronized control for producing stereophonic effects.

C. J. Carter and Pye Ltd. Application date 23rd February, 1943.

584 457.—Synchronizing system for multichannel communication on time-modulated pulsed signals.

Standard Telephones and Cables Ltd. (assignees of E. M. Deloraine and N. H. Young). Convention date (U.S.A.) 29th September, 1943.

584 513.—Two-way short-wave set on which a fraction of the transmitted wave is used to heterodyne the incoming signal through an adjustable aerial coupling.

" Patelhold " Patentverwertungs etc. A.G. Convention date (Switzerland) 26th July, 1943.

584 726.—Combined pulse transmitter and receiver, in which a radio-frequency circuit and a quench-frequency circuit are both operated by the same valve.

Marconi's W.T. Co. Ltd., N. M. Rust, G. E. Partington, D. L. Plaistowe and D. J. Fewings. Application date 4th February, 1941.

584 729.—Multichannel signalling system utilizing variable-width and fixed-amplitude pulses, with " stroboscopic " switching or synchronizing pulses.

W. A. S. Butement and A. J. H. Oxford. Application date 4th December, 1942.

584 733.—Variable-width pulsed signalling system, on which calling or supervisory signals are superposed by the deliberate suppression of selected pulses.

J. G. Macmillan and Z. Jelonek. Application date 7th September, 1943.

SUBSIDIARY APPARATUS AND MATERIALS

584 072.—Piezo-electric crystal resiliently mounted at the opposite ends of a nodal line of oscillation.

The British Thomson-Houston Co. Ltd. (communicated by the General Electric Co.). Application date 27th October, 1944.

584 088.—Piezo-electric crystal with a mounting which is adjustable, resilient and free from damping.

Standard Telephones and Cables Ltd. (assignees of S. A. Bokovoy and K. K. Garrison). Convention date (U.S.A.) 6th December, 1943.

584 315.—Glass-wool cover or flexible envelope for holding a thermionic valve in its socket.

D. R. Gamlen and J. Buckingham. Application date 2nd May, 1944.

584 329.—Multivibrator type of circuit in which a square-shaped wave triggers a cycle of operations through a step-by-step reduction of the anode voltage, suitable for counting operations or as a frequency-divider.

F. C. Williams. Application date 1st September, 1944.